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RANGE/RANGE RATE SUBSYSTEM (RRS) FOR COMPLETELY INTEGRATED REFE--ETC(U)
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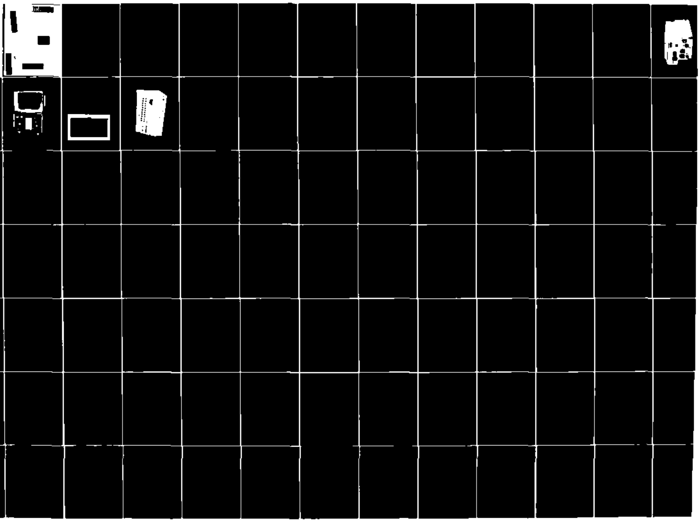

RANGE/RANGE RATE SUBSYSTEM (RRS)

FOR

COMPLETELY INTEGRATED REFERENCE INSTRUMENTATION SYSTEM (CIRIS).

CUBIC CORPORATION

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FOREWORD

This Technical Report was produced by the Cubic Corporation, San Diego, California in compliance with Data Item A005 of the Contract Data Requirements List, USAF Contract F29601-71-C-0079. The report covers the work performed by the Cubic Corporation in connection with the design, development, fabrication, test, and delivery to the Air Force of a Range/Range Rate Subsystem for use in the Completely Integrated Reference Instrumentation System (CIRIS).

The Guidance Test Division, Air Force Special Weapons Center (AFSWC), Holloman AFB, New Mexico, was the contracting agency for the Air Force. Project direction and administrative assistance were provided by Mr. Peter U. Zagone, Chief of Operation Test Branch, and Captain Melvin Birnbaum, USAF, CIRIS Technical Director for the Guidance Test Division.

The contractor's identification number assigned to this Final Technical Report is Cubic Document No. FTR/526-1.

This technical report has been reviewed and approved.

ABSTRACT

An equipment procurement comprising a Range/Range Rate Subsystem (RRS) for the Completely Integrated Reference Instrumentation System (CIRIS) is described in detail in a final technical report. An RRS composed of an airborne interrogator and four ground-based transponders is designed, fabricated, factory acceptance-tested, and delivered to the Air Force. With the interrogator operating under the control of the CIRIS airborne computer, and with ground transponders set up for operation at known (surveyed) locations, the RRS measures slant ranges from 200 feet to 200 miles with instrumental accuracies to within 3 feet rms, and measures range rates varying from -5000 to +5000 fps with instrumental accuracies to within 0.03 fps rms under range acceleration conditions from -1000 to +1000 fps^2. The CIRIS computer employs the RRS data to optimally update an inertial measurement unit, providing an accurate, real-time aircraft position, velocity and attitude reference system for aircraft flight test programs. Predesign analyses of subsystem requirements are discussed, delivered equipment units are pictured and described, and details of the computer/interrogator interface are presented. New circuit developments include a microstripline preamplifier, fully limiting receiver and 7.5W power amplifier, all operating at 1.6 GHz. A transponder verification problem is identified and its solution is provided. The report concludes that the RRS units will reliably meet all specified performance criteria.

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SECTION I

INTRODUCTION AND SUMMARY

1. INTRODUCTION.

a. Purpose of Report. This report describes the predesign analyses, design work, fabrication, and postproduction tests conducted in the development, production, and delivery to the Air Force of a Range/Range Rate Subsystem (RRS) for use in the Completely Integrated Reference Instrumentation System (CIRIS). The RRS equipment covered in this report is configured to meet the requirements specified in Purchase Request FY617-71-16014 initiated by the Guidance Test Division, Air Force Special Weapons Center (AFSWC), Holloman AFB, New Mexico.

b. Purpose of Range/Range Rate Subsystem. The RRS consists of an airborne interrogator and ground-based transponders designed so that, when the interrogator is interfaced with an airborne digital computer and the transponders are installed at surveyed locations on the ground, the subsystem performs the following functions:

(1) It provides an air-to-ground-to-air (AGA) data link for activating any transponder selected by the computer, and for verifying that the selected transponder responds to the interrogation.

(2) It provides an AGA coherent carrier tracking loop for deriving range rate data accurate to within 0.03 fps (rms), measuring range rates varying from -5000 to +5000 fps with range accelerations from -1000 to +1000 fps^2.

(3) It provides an AGA range modulation loop for deriving range data accurate to within 3 feet (rms), measuring line-of-sight ranges from 200 feet to 200 miles.

(4) It stores the range, range rate, and data quality data for sequential readout by the computer.

The CIRIS digital computer uses valid range and range rate data supplied by the RRS to optimally update an Inertial Measurement Unit (IMU) to provide a real-time aircraft position, velocity and attitude reference system for use in aircraft flight test programs.

c. Program Objectives. The program objectives included the following:

(1) To design, fabricate and test an airborne interrogator meeting the requirements of Addendum One to the Statement of Work (Specifications, Interrogator, Range/Range Rate).

(2) To design, fabricate and test four ground transponders meeting the requirements of Addendum Three to the Statement of Work (Specifications, Transponder, Range/Range Rate).

(3) To provide a ground antenna for each transponder configured for optimum coverage and designed to collapse into a package small enough for easy transport to a new site.

(4) To provide an aircraft antenna that is physically and electrically compatible with existing AFSWC guidance test aircraft.

(5) To design, fabricate, and provide an Interrogator Test Set configured to simulate computer commands and display RRS output data.

(6) To analyze the reliability, maintainability, and thermal dissipation aspects of the equipment design, and furnish all data items specified in the Purchase Order.

(7) To generate procedures and conduct acceptance tests as specified in Addendum Four to the Statement of Work (Specifications, Radio Ranging Subsystems).

2. SUMMARY OF CONTRACT PERFORMANCE.

a. RRS Equipment Units Delivered. The equipment units that were designed, built, acceptance-tested, and delivered to the Air Force for integration into the CIRIS program consisted of one Interrogator RT-1091/URQ-22, associated aircraft antenna, and interconnecting cables; four Transponders RT-1091/URQ-22, associated antennas, cables, and batteries; and one Indicator-Interrogator Set Test Set TS-3302/URQ-22.

(1) Interrogator. The airborne interrogator (figure 1) is packaged to fit a standard 3/4 ATR short case for quick, easy installation and removal. This unit is 8 in. high, 8 in. wide and 13 in. deep, and weighs 30 lb. Some of the principal electrical characteristics are listed below.

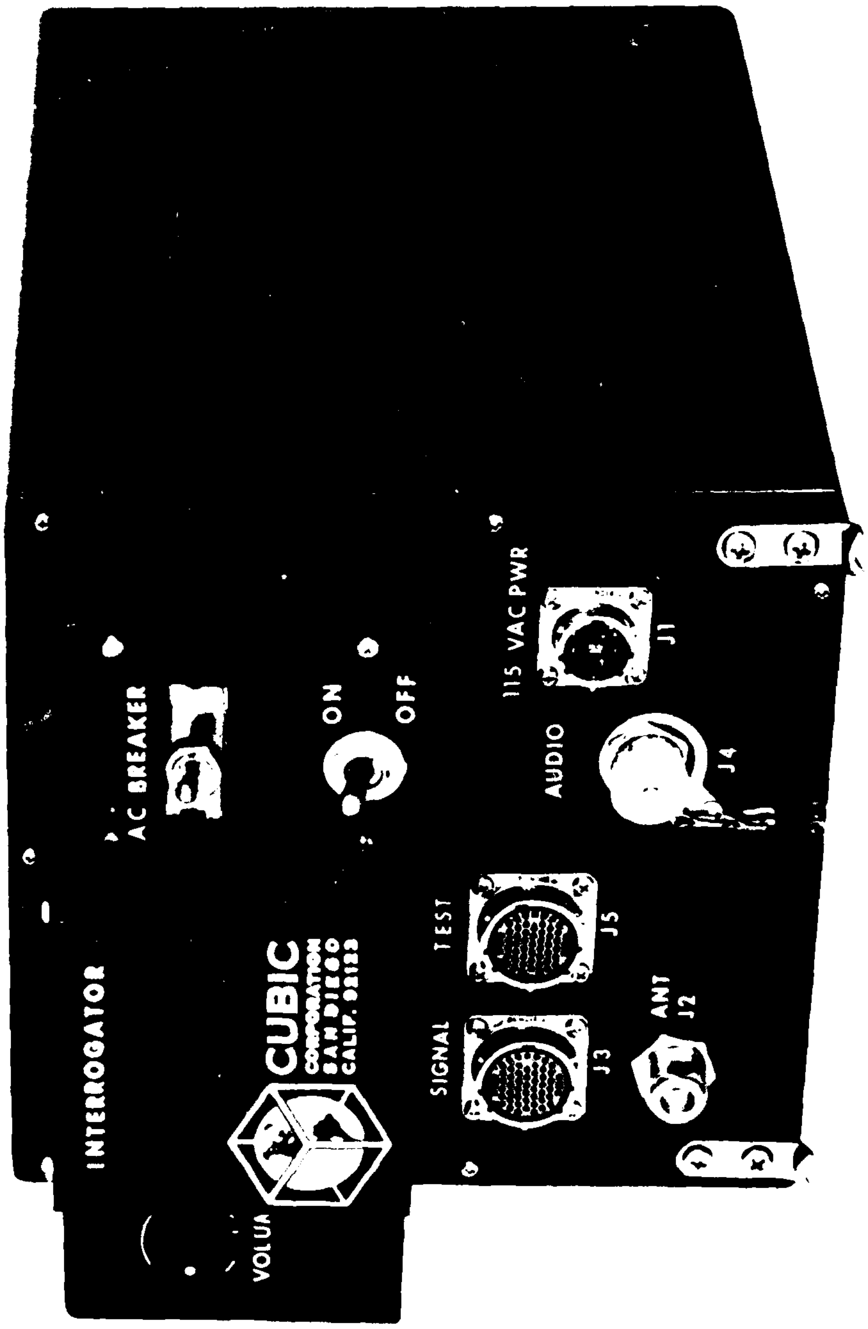
(a) Operating frequencies: 1630 MHz transmit, 1564.8 MHz receive

(b) Transmit power: 3.5W minimum

(c) Receiver sensitivity: -110 dBm

(d) Data output to computer: two 20-bit range rate words, four 11-bit range words, one 7-bit data quality word

(e) Input power required: 115/200 Vac, 3-phase, 400-Hz, 120W peak


INTERROGATOR
CUBIC
CORPORATION
SAN DIEGO
AC BREAKER
ON
OFF
115 VAC PWR
J1
AUDIO
J4
TEST
J5
SIGNAL
J3
ANT
J2

(2) Airborne Antenna. The aircraft antenna (figure 2) is a quarter-wave stub housed within an aerodynamically shaped radome and configured for easy mounting onto the belly of the flight test aircraft. The antenna illuminates the lower hemisphere below the aircraft, providing 360 degrees coverage in azimuth. Characteristic impedance is 50 ohms, and maximum VSWR is 1.5;1.

(3) Transponder. Each transponder (figure 3) is a rugged, man-transportable, ground-based unit capable of sustained unattended operation when emplaced at a surveyed location. The portable carrying case that houses the transponder electronics also accommodates a 7-ampere-hour rechargeable battery, an ac power supply, and battery charger. The cover contains a hinged lid that forms a compartment for the power and antenna cables and a voice headset. Some of the principal characteristics are listed below.

(a) Operating frequencies: 1564.8 MHz transmit, 1630 MHz receive

(b) Transmit power: 3.5W minimum

(c) Receiver sensitivity: -110 dBm

(d) Size: 16 in. wide, 10 in. high, 14 in. deep

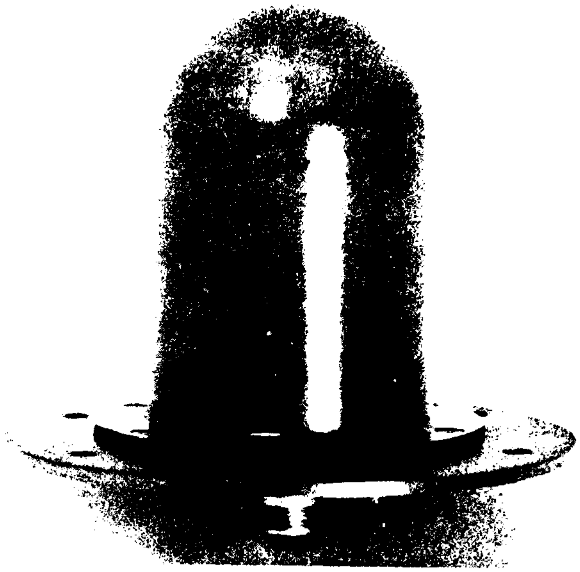
(e) Weight:

(1)	Transponder	13 lb.
(2)	Power Supply Charger plus case	16 lb.
(3)	Battery	13 lb.
	Total	42 lb.

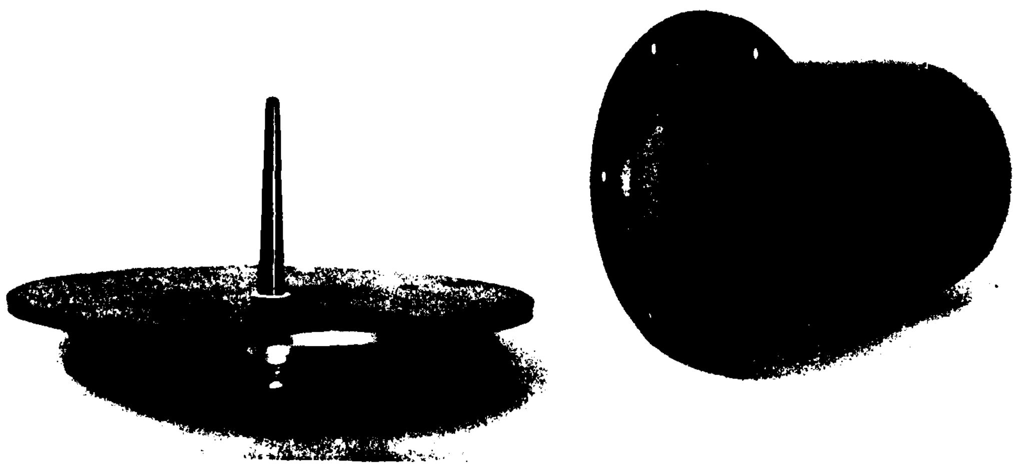
(f) Input power required: 28 Vdc, 6W standby power, 50W transmit. Sources: 28V battery or 115V, 60-400 Hz single-phase ac

(4) Ground Antenna. The ground antenna (figure 4) consists of a discone radiating element located in the center of an aluminum-framed ground plane 4 feet square. The radiating element is protected by a fiberglass/epoxy radome, and both radome and antenna connector are sealed against moisture. This antenna assembly folds up to form a 2- by 4-foot unit with carrying handle. Like the airborne quarter-wave stub, the discone antenna element provides 360° coverage in azimuth, and offers a 50-ohm impedance and VSWR of 1.5:1.

(5) Interrogator Test Set. The interrogator test set (figure 5) is a special-purpose test instrument designed specifically to (a) provide the control features normally performed by the computer and (b) provide the readout features that enable



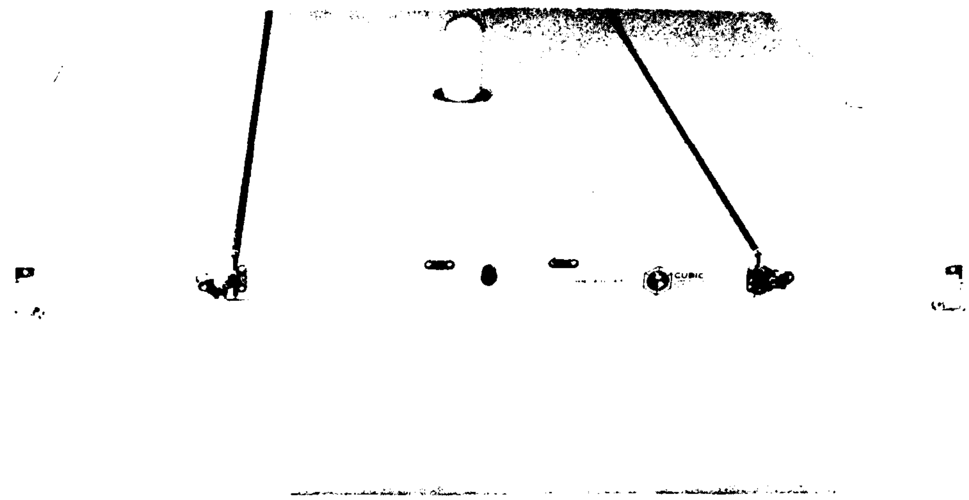
A. RADOME INSTALLED



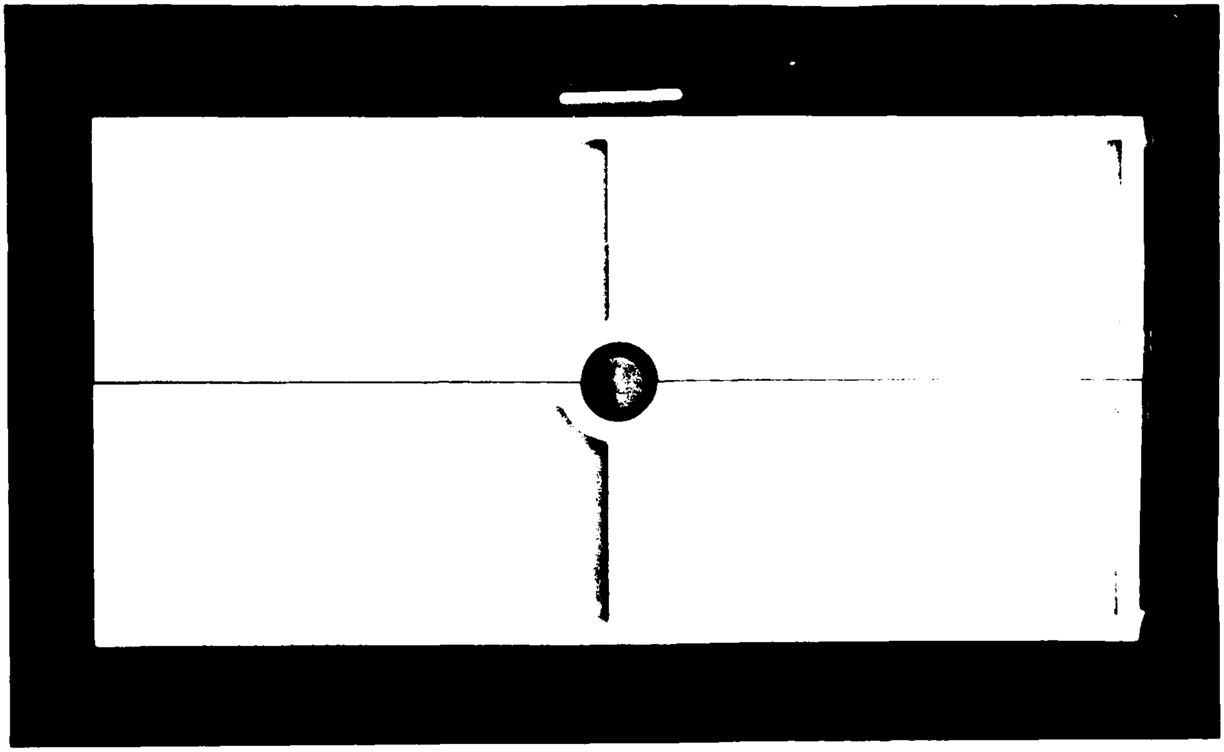
B. RADOME REMOVED

Figure 2. Interrogator Antenna (P/N 128275-1)

Figure 3. Transponder RT-1092/URQ-22



A. UNFOLDED FOR OPERATION



B. FOLDED FOR TRANSPORT

Figure 4. Transponder Antenna (P/N 128285-1)

Figure 5. Indicator-Interrogator Set Test Set TS-3301/URQ-22

the performance of various interrogator circuits to be checked and evaluated. When the interrogator is rf-linked to a transponder, the test set permits the operation of the overall RRS to be checked and evaluated.

b. RRS Data Items Submitted. Data end items delivered in compliance with the Purchase Order included the following:

(1) Acceptance Test Procedure, Cubic Document ATP/526-1

(2) Acceptance Test Report, Cubic Document TR/526-1

(3) Preliminary Instruction Manual, Cubic Document PHB/526-1

(4) Reliability and Maintainability Analysis, Cubic Document RMR/526-1

(5) System Safety Program Plan, Cubic Document SSPP/526-1.

3. ORGANIZATION OF REPORT. Section II of this report presents the results of the predesign analyses conducted to ensure that the RRS equipment would meet or better all performance specifications. Section III describes the RRS equipment supplied in compliance with the Purchase Order, and includes a discussion of the two antenna types. Section IV presents a detailed, comprehensive description of the RRS/Computer interface arrangements. Section V examines some of the new circuit design features developed for the RRS, and section VI discusses the equipment packaging techniques employed in fabricating the RRS unit. Section VII discusses reliability and maintainability, and section VIII presents the conclusions and recommendations.

SECTION II

PREDESIGN ANALYSIS OF SYSTEM PARAMETERS

1. INTRODUCTION. This section presents the results of the predesign analyses conducted in connection with the development of the RRS equipment described in section III.

2. SIGNAL-TO-NOISE RATIO (SNR) AND ERROR SUMMARIES. The following summarizes the significant SNR and error values derived from the detailed calculations and analyses provided in subsequent paragraphs of this section. The transmitter output powers, carrier loop bandwidths, and noise figures indicated apply to both interrogator and transponder inasmuch as the electrical designs of the rf subsystem for both components are nearly identical. It is significant to note that the carrier loop has the smallest safety margin and hence represents the weakest link in the subsystem. Accordingly, if the carrier signal is acquired, then data accuracy is assured.

a. Received Signal Power (Maximum Range = 200 Mi.)

+36 dBm	4-watt transmitter output power
-1 dB	Antenna cable, transmit path (12 ft RG/217 which exhibits
+35 dBm	8 dB loss per 100 ft at 1600 MHz)
+3 dB	Antenna gain (ground antenna)
+38 dBm	
-147 dB	Path attenuation, 200 mi.
-109 dBm	
0 dB	Antenna gain (aircraft antenna)
-109 dBm	
-1 dB	Antenna cable, receive path (12 ft RG/217)
-110 dBm	Signal power at receiver input

b. Carrier Loop SNR at 200 Mi.

-174 dBm	KT
7 dB	Noise figure
62 dB	Noise bandwidth (1.5 MHz in IF)
-105 dBm	Noise power
-110 dBm	Received signal power
-5 dB	
1 dB	Loss due to limiter
1 dB	Carrier loss due to index (0.8)
-7 dB	Carrier-to-noise ratio in IF
+24 dB	Bandwidth exchange to 2 B_L = 6 kHz
+17 dB	SNR in carrier loop
-5 dB	Data loop noise where 2 (BW_D) = 12 kHz
+12 dB	SNR carrier loop with full modulation at 200 mi.

c. Range Rate Data SNR at 200 Mi.

+12 dB	SNR in carrier loop
+54 dB	SNR improvement due to vclo (÷ 128) and BW (200 Hz)
+66 dB	SNR range rate servo
+34 dB	SNR required for 1 σ of 0.03 ft/sec
+32 dB	Safety margin in range rate servo

d. Range Data SNR at 200 Mi.

+12 dB	SNR in carrier loop (B_L = 3 kHz)
+26 dB	β^2 FM improvement (M = 20)
+19 dB	Bandwidth exchange to B_L = 40 Hz
+57 dB	Range data SNR
38 dB	SNR required for 3 ft rms jitter
+19 dB	Safety margin in range data

e. Carrier Acquisition Link SNR.

-7 dB	Carrier-to-noise ratio in IF (index = 0.86)
-1 dB	Carrier loss (index = 1)
-8 dB	Carrier-to-noise ratio in IF
+25 dB	Bandwidth improvement to B_L = 2.5 kHz
+17 dB	Acquisition link SNR
+12 dB	SNR required for 10^{-5} probability of error
5 dB	Margin above 10^{-5} error probability

f. Range Data Error Summary.

Error Source	RMS Value (ft)
Oscillator stability (1 ppm)	0.33
Interrogator calibration	0.33
Transponder calibration	0.33
Drift (transponder)	0.33
(interrogator)	0.33
Velocity (max. v)	0.02
Acceleration (max. a)	0.20
Noise (external)	0.33
Noise (equipment)	0.17
Resolution (1-bit uncertainty)	0.33
Range reading ref "0" crossing (max. v)	0.45
Multipath	2.00
RSS	2.24

g. Range Rate Data Error Summary.

Error Source	RMS Value (ft/sec)
Oscillator stability (1 ppm)	0.002
Acceleration (max. a)	0.010
Noise (external)	0.002
Noise (equipment)	0.003
Resolution (1 bit)	0.007
Multipath	0.010
RSS	0.016

3. PERFORMANCE AND ERROR ANALYSIS FOR COHERENT-CARRIER RRS. This paragraph describes the desired performance characteristics, the required parameters, and the predicted errors for the full-limiting, coherent-carrier, range and range-rate measuring subsystem described in section III. The analysis centers on the interrogator since, with its range and range-rate servo loops, it represents the worst-case condition. However, transponder factors affecting the subsystem are shown wherever relevant.

a. Coherent Carrier Loop.

(1) Operating Parameters (Interrogator).

Frequencies:	Received carrier	1564.8 MHz
	First if.	70.6 MHz
	Second if.	16.3 MHz
	Transmitted carrier	1630.0 MHz
Dynamic Range	75 dB (-110 dBm to -35 dBm)	
Range Modulation	Four tones in 250-kHz band	
	Indexes (maximum):	20 for 1 tone, 2 for remaining 3 tones
	Modulation buildup:	linear ramp, ≈ 10 ms duration
	Data tone loop gain:	≈ 30 dB
	Open loop bandwidth, data filter	≈ 50 Hz × 4 = 200 Hz
	Data noise bandwidth:	2 (200) = 400 Hz

Drifts:	Doppler:	±16 kHz max.
	Transmitter (Temp. Comp. Xtal Osc.):	±1 ppm
	First LO (Temp. Comp. VCO):	±10 ppm
	Acceleration:	±3.2 kHz/sec
Carrier Acquisition Time:		25 ms max.

(2) Minimum IF Bandwidth. Under closed data loop operation the if. index is:

$$\text{Index/feedback for 1 tone} = \frac{20}{31.6} = 0.64$$

$$\text{for 3 tones} = 3(2/31.6) = 0.19$$

$$\text{Peak deviation} = 0.64 + 0.19 = 0.83$$

and the carrier if. bandwidth B_{if} required by the complex modulating signal is given by:

$$B_{if} = 2\ (\Delta f_m) + \text{drifts}$$

where,
Δf = peak frequency deviation

f_m = highest baseband frequency

$$B_{if} = 2\left[\ 0.83\ (250) + 2\ (250)\ \right] + \text{drifts}$$

$$B_{if} = 1.440\ \text{MHz} + \text{drifts}$$

The drifts affecting the if. bandwidth are calculated as follows:

Factor	Interrogator	Transponder
Doppler	±16.0 kHz	±8.3 kHz
First LO (10 ppm)	±15.6 kHz	±16.3 kHz
Transmitter (1 ppm)	±1.6 kHz	±1.6 kHz
Totals	33.2 kHz	26.2 kHz

therefore,

$$B_{if} = 1.440\ \text{MHz} + 0.033\ \text{MHz} = 1.473\ \text{MHz}$$

setting the design requirement at

$$B_{if} = 1.5\ \text{MHz}$$

Temperature-compensated voltage-controlled crystal oscillators (first LO above) suitable for aerospace and portable equipment applications are available which are sufficiently stable to permit long-term unattended operation (up to a year or longer) without readjustment of center frequency.

(3) <u>Loop Natural Frequency and Loop Bandwidth</u>. The maximum acceleration will cause a frequency rate-of-change ($\Delta\dot{\omega}$) at the interrogator of

$$\Delta\dot{f} = \pm 2 \frac{a_{max}}{c} f_{TS} = \text{Hz/sec}$$

then,

$$\Delta\omega = \pm 2 \; \frac{1000 \; (1564.8 \times 10^6) \; (2\pi)}{9.833 \times 10^8} \approx 20,000 \text{ radians/sec}^2$$

where

a_{max} is the maximum acceleration, and

f_{TS} is the transponder center frequency under static (non-moving) conditions

The loop natural frequency is governed in this case by the allotted 25-ms maximum sweep acquisition time, system drift and doppler of 34 kHz. The rate of change of the sweep frequency Δf therefore is

$$\Delta\dot{f} = \frac{2(3.32 \times 10^4 \text{ Hz})}{2.5 \times 10^{-2}\text{sec}} = 2.66 \times 10^6 \text{ Hz/sec}$$

$$\text{or } \Delta\dot{\omega} = 1.67 \times 10^7 \text{ radians/sec}^2$$

To insure greater than 99% probability of lock during the first sweep, a sweep rate of $\Delta\dot{\omega} = \frac{\omega_n^2}{2}$ (1/2 the theoretical limit) is reasonable,

therefore $\omega_n^2 = 2(1.67 \times 10^7)$

and $\underline{\omega_n = 5.78 \times 10^3 \text{ radians}}$

The loop bandwidth B_L (in Hz) as a function of ω_n and the loop damping factor ζ is expressed by:

$$B_L = \frac{\omega_n}{2} \left(\zeta + \frac{1}{4\zeta}\right)$$

At system threshold, ζ is set at 0.707, therefore

$$B_L = \frac{5.78 \times 10^3}{2} \left(0.707 + \frac{1}{4(.707)}\right) = 3.07 \text{ kHz}$$

setting the design requirement at $\underline{B_L = 3 \text{ kHz}}$

Without VCO sweeping, the maximum pull-in time (T_p) is proportional to the frequency offset and loop bandwidth. For a damping factor ζ of 0.707,

$$T_p \approx 4.2 \frac{\Delta f^2}{B_L^3} \text{ sec} = 4.2 \frac{(3.32 \times 10^4)^2}{(3 \times 10^3)^3}$$

$$T_p \approx 0.17 \text{ sec}$$

This 170-ms acquisition time is of course excessive, and shows that the system design requires VCO sweeping as an acquisition aid.

(4) Acceleration Error. The range rate error in a 1-second measurement will reach the quantizing limit when the phase error $\theta_a = \frac{2\pi}{16} = 0.393$ radians.

The maximum error in the receiver carrier loop is

$$\theta_a = \frac{\Delta\dot{\omega}}{\omega_n^2} = 0.0006 \text{ radians}$$

where

$$\Delta\dot{\omega} = 20,000 \text{ radians/sec}^2 \text{ at } a_{max} = 1000 \text{ ft/sec}^2$$

and

$$\omega_n = 5.78 \times 10^3 \text{ radians}$$

This results in an insignificant error in the range rate of

$$\frac{0.0006}{0.393} \times 0.0196 \text{ ft/sec} = \underline{0.00003 \text{ ft/sec}}$$

(5) Velocity Error. The minimum required loop gain K_V is determined by the static phase error allowed under worst-case drifts,

$$\text{thus, } K_V = \frac{\Delta\omega}{\theta} = 2.14 \times 10^6 = \underline{127 \text{ dB}}$$

where

θ = static phase error of 0.1 radian

and

$\Delta\omega = 34 \times 10^3 (2\pi) = 2.14 \times 10^5$ radians/sec

Therefore the minimum loop gain needed at threshold is 127 dB.

(6) Noise Figure.

(a) The receiver front-end noise figure, NF is

$$NF = Nf_1 + \frac{Nf_2 - 1}{G_1} + \frac{Nf_3 - 1}{G_1 G_2}$$

where

Nf_1 = 2.5 (4-dB noise figure of first stage)

Nf_2 = 2.5 (4-dB noise figure of second stage)

Nf_3 = 16 (12-dB noise figure of mixer and first if. stage)

$G_1 = G_2$ = 6 (8-dB gain in each of 2 rf stages)

therefore,

$$NF = 2.5 + \frac{1.5}{6} + \frac{15}{36} = 3.17 = 5.0 \text{ dB}$$

(b) Filter and Circulator loss = 1.5 dB

(c) Noise figure total = 5 dB + 1.5 dB = 6.5 dB

For RRS design calculations, use NF = 7 dB.

(7) Receiver SNR.

(a) The carrier-to-noise ratio in the receiver if. is

-174 dBm	KT
62 dB	IF noise BW of B_{if} = 1.5 MHz
-112 dBm	

-112 dBm	Ideal noise power
7 dB	Noise figure
-105 dBm	Actual noise power
-110 dBm	Received signal power
-5 dB	
-1 dB	SNR loss due to limiter
-6 dB	
-1 dB	Carrier loss due to index (0.83) in if.
-7 dB	Carrier-to-noise in if.

(b) The data loop effects are as follows: the four tone filters, total open-loop bandwidth is 4 (50) = 200 Hz, undergo a closed loop bandwidth incr by the feedback factor. This yields a closed-loop data bandwidth of BW_D = 200 (≈ 6 kHz.

(c) The carrier loop SNR at maximum range is:

-7 dB	Carrier-to-noise in if. at B_{if} = 1.5 MHz
+24 dB	Bandwidth exchange to 2 B_L = 6 kHz
+17 dB	
-5 dB	Data loop noise at 2 (BW_D) = 12 kHz
+12 dB	SNR in carrier loop (system operation at thres with full modulation).

b. Range Rate Data Servo Loop.

(1) Dynamics. The maximum doppler and doppler rate as seen by range-rate servo input is the carrier rate divided by 128.

$$d_{max} = \frac{\pm 16 \text{ kHz}}{128} = \pm 125 \text{ Hz}$$

$$a_{max} = \frac{\pm 3200 \text{ kHz/sec}}{128} = \pm 25 \text{ Hz/sec}$$

(2) $\dot{R}$ Servo Loop Bandwidth. Since sufficient signal-to-noise rati in the system, a maximum upper tracking error of 0.01 ft/sec is assigned to rang rate. This is less than one-half of the system resolution. The tracking error expressed in phase θ_a is:

$$\theta_{a/\Delta T} = 0.01 \text{ ft/sec} \times \frac{1 \text{ bit/sec}}{0.02 \text{ ft/sec}} \times \frac{2\pi \text{ radians/sec}}{2048 \text{ bits/sec}}$$

$$\theta_{a/\Delta T} = 0.00154 \text{ radians/sec}$$

The loop natural frequency ω_n can be calculated:

$$\omega_n^2 = \frac{\Delta\dot{\omega}}{\theta_a}$$

where

$$\Delta\dot{\omega} = a_{max} = +25\ (2\pi)\ \text{rad/sec}^2 = 157\ \text{rad/sec}^2$$

Therefore,

$$\omega_n = \left(\frac{\Delta\dot{\omega}}{\theta_a}\right)^{1/2} = \left(\frac{157}{0.00154}\right)^{1/2} = 320\ \text{radians}$$

Setting the damping factor (ζ) to 0.707, the loop bandwidth is

$$B_L = \frac{\omega_n}{2}\ (\zeta + \frac{1}{4}\zeta)\ \text{Hz} = 170\ \text{Hz}$$

which sets the design parameter at $\underline{B_L = 200\ \text{Hz}}$

(3) Lock-Up Time and Settling Time. The maximum lock-up time is 5 ms. The settling time (t) is obtained from:

$$\frac{1}{2048} = \exp\ (-t\ \omega_n/2\zeta)$$

where

$$t = 0.0036\ (7.6) = 28\ \text{ms}.$$

Thus, lock-up plus settling time = 33 ms.

The "settling" time delay for the range rate lock control and $\dot{R}$ lock data quality bit is set at 30 ms after the $\dot{R}$ servo loop locks.

The center frequency of the servo VCO is 10.866 MHz, corresponding to f_o. The VCO must exhibit a tracking capability at maximum velocity of

$$\pm 125\ \text{Hz} \times 2048 = \pm 256\ \text{kHz}$$

(4) Range Rate SNR. The signal-to-noise ratio in the carrier loop is +12 dB at a noise bandwidth of 6 kHz. Since the input to the data servo comes from the VCLO in the carrier receiver, the signal input to the servo is 128 times better; i.e., 12 dB + 42 dB = +54 dB in a 6 kHz noise bandwidth. The improvement provided by the 400-Hz noise bandwidth of the range rate servo raises the range rate servo loop SNR to 54 + 12 = +66 dB.

The required SNR in the range rate servo for $1\sigma = 0.03$ ft/sec error is $\theta = 3(0.00154) = 0.00462$ radians. Since an SNR of $\frac{1}{2\sigma^2} = \frac{1}{2(0.00462)^2} = 2340 = 34$ dB is required, the safety factor is $66 - 34 = 32$ dB.

(5) External Noise Error. The actual phase jitter due to external noise is:

$$\sigma = \frac{1}{\sqrt{2\ S/N}} = \frac{1}{\sqrt{2\ (4)\ 10^6}} = 3.53 \times 10^{-4} \text{ radians}$$

converting this phase error to feet over a 1-second count period gives

$$1\sigma = 3.53 \times 10^{-4} \text{ rad/sec} \times \frac{1 \text{ ft/sec}}{0.154 \text{ rad/sec}}$$

or,

$$1\sigma = 0.0023 \text{ ft/sec}$$

(6) Range Rate Resolution. The range rate resolution is expressed in feet-per-bit, and is a function of carrier frequency, velocity of propagation of the electromagnetic wave, and the length of time of the measurement period, making the measurement period the only variable in the system. $\dot{R}$ is obtained from the formula for the doppler offset of the received carrier frequency of the interrogator plus the multiplication of the doppler offset in the interrogator, thus:

$$d = 16 \left[+2 \frac{v}{c} f_{TS} + \left(\frac{v}{c} \right)^2 f_{TS} \right]$$

where:

d = interrogator received frequency offset (multiplied)

v = relative differential velocity between interrogator and transponder

c = velocity of propagation of electromagnetic wave

f_{TS} = center frequency of transponder (static conditions)

The constant 2 results from the two-way coherent link. The constant 16 results from the times-16 multiplication of the doppler frequency portion after it is received.

The velocity of propagation of electromagnetic waves in a vacuum is officially taken by the International Union of Geodesy and Geophysics as

$$c_0 = 299,792.5 \pm 0.3 \text{ km/sec}$$

Since the RRS measurement is expressed in the English system, the coversion constant of 3.2808399 feet per meter is used. Therefore,

$$c_o = 9.835712 \times 10^8 \text{ ft/sec}$$

and in a medium other than a vacuum,

$$c = \frac{c_o}{n}$$

where n is the index of refraction.

Experience with many DME systems show that, in applying corrections to the velocity of propagation in the atmosphere, a correction of N = 320 ppm is a good mean value to use for most applications. Thus,

$$c = \frac{c_o}{1 + \frac{320}{10^6}} = 9.832565 \times 10^8 \text{ ft/sec}$$

The transponder carrier frequency is chosen to be a coherent ratio of $\frac{n}{n+1} = \frac{24}{25}$ of the selected interrogator frequency of 1630.000 MHz, or,

$$f_{TS} = \frac{24}{25}(1630) = 1564.800 \text{ MHz}$$

The value of the second-order term in the doppler equation is insignificant at maximum velocity and can be ignored; so that,

$$d = 16\left(\frac{v}{c}\right)^2 f_{TS} = 0.647 \text{ Hz}$$

Using the above terms, the range rate resolution is calculated in feet-per-bit as follows:

$$\text{Resolution} = \frac{v}{d} = \frac{1}{2(16)} \frac{c}{f_{TS}} \frac{\text{ft/sec}}{\text{cycles/sec}} \times \frac{1 \text{ cycle}}{\text{bit}}$$

$$= \frac{9.832565 \times 10^8}{2(16)1564.8 \times 10^6} \frac{\text{ft}}{\text{bit}} = 0.019636 \frac{\text{ft}}{\text{bit}}$$

Using a 1-second count period, the range rate resolution would therefore be 0.019636 ft/sec.

c. Range Data Servo.

(1) Range Resolution. Since the system range resolution is specified at 1 foot, the frequency of the range master oscillator becomes a function of (a) this resolution, (b) the range servo counter constant, and (c) the velocity of propagation corrected to 320 ppm. With a servo counter constant of 2048, each cycle of the fine modulation is divided by that figure. Also, because the actual range measured is the round-trip or two-way slant distance, the required modulation frequency becomes:

$$f_M = \frac{c}{\lambda} = \frac{9.832565 \times 10^8 \text{ ft/sec}}{(2)\ (2048) \text{ ft/cycle}}$$

$$f_M = 240.0528 \text{ kHz}$$

Since stable crystal oscillators are easier to obtain at frequencies above 1 MHz, the frequency f_M is actually derived from a temperature compensated crystal oscillator at 8 times f_M or $f_{M_{osc}} = 8 f_M = 1.920423$ MHz

(2) Dynamics. The maximum doppler and doppler rates as seen by the range servo input is

$$d_{max} = \pm 2 \frac{v}{c} f = 2.8 \text{ Hz}$$

where

v = maximum velocity (5000 ft/sec)

c = velocity of propagation (9.8×10^8 ft/sec)

f = highest modulating frequency (270 kHz)

$$a_{max} = \pm 2 \frac{v/c_{max}}{c} f = 0.56 \text{ Hz/sec}$$

(3) Required Loop Bandwidth. The loop bandwidth must be chosen to keep the peak-to-peak (3σ) error less than 1 foot at maximum acceleration, or

$$\theta_a = 0.5 \text{ ft} \times \frac{2\pi}{2048} = 0.00154 \text{ radians (maximum)}$$

The loop natural frequency ω_n is calculated from:

$$\omega_n^2 = \frac{\Delta\dot{\omega}}{\theta_a} = \frac{0.56\ (2\pi)}{0.00154}$$

$$\omega_n = 48 \text{ radians}$$

Setting the damping factor ζ to 0.707,

$$B_L = \frac{\omega_n}{2} \quad \zeta + \frac{1}{4\zeta} \quad \text{Hz} = 20 \text{ Hz minimum}$$

Examining the settling time for this loop bandwidth:

$$\frac{1}{2048} = \exp\left(-t\omega_n / 2\zeta\right)$$

$$t = 7.6\ (0.0295) = 0.224 \text{ sec}$$

(4) SNR and External Noise Error. From an operational viewpoint it would be better to reduce the settling time by opening the bandwidth of the servo if sufficient signal-to-noise exists so as not to degrade the measurable accuracy. The data output SNR in the receiver is improved by the square of the index β. For the fine tone, $\beta = 20$ and $\beta^2 = 26$ dB. For the intermediate, coarse and very coarse tones, $\beta = 2$ and $\beta^2 = 6$ dB. Thus:

$$\text{SNR (fine tone)} = +12 \text{ dB} + 26 \text{ dB} = 38 \text{ dB}$$

$$\text{SNR (other 3 tones)} = +12 \text{ dB} + 6 \text{ dB} = 18 \text{ dB each}$$

A peak-to-peak or 3σ error of 1 foot due to noise in the servo word required a 1σ error of

$$\theta = 0.33 \text{ ft external noise}$$

$$\theta = 0.33 \text{ ft } \frac{2\pi}{2048} = 0.001 \text{ radians}$$

therefore,

$$\text{Required SNR} = \frac{1}{2\sigma^2} = \frac{1}{2(10)^{-6}} = 5 \times 10^5 = 57 \text{ dB}$$

Setting this SNR requirement of 57 dB on the fine servo at threshold requires a bandwidth improvement of $57 - 38 = 19$ dB = 80. Therefore the fine servo loop bandwidth requirement is $B_L = \frac{3 \text{ kHz}}{80} = 38$ Hz

or, as the design parameter, set the FN servo to: $B_L = 40$ Hz

The other three servos (INT, CS, and VS) are also set at $B_L = 40$ Hz for convenience and interchangeability. The SNR of these loops at system threshold is:

$$18 \text{ dB} + 19 \text{ dB} = 37 \text{ dB} = 5000$$

The phase error therefore becomes

$$1\sigma = \frac{1}{\sqrt{2\ S/N}} = \frac{1}{\sqrt{(10^4)}^{1/2}} = 0.01 \text{ radians}$$

or

$$\text{p-p } 3\sigma = 0.03 \text{ radians}$$

This corresponds to a peak-to-peak jitter of

$$0.03 \text{ rad}/0.003 \frac{\text{rad}}{\text{bit}} = 10 \text{ bits}$$

Since the ambiguity resolution algorithm corrects up to 128 bits, a 13-to-1 safety factor exists at threshold in this respect.

(5) Lock-Up Time and Settling Time. The lock-up time is between 10 and 20 ms. The settling time is

$$\frac{1}{2048} = \exp\left(-t\omega_n/2\zeta\right)$$

where

$$\omega_n = 76$$

thus,

$$\underline{t = 7.6\ (.0187) = 142 \text{ ms}}.$$

The composite range lock delay control for data quality is set for 150 ms, requiring all servos to be locked for this length of time prior to being sampled at the end of the interrogation cycle.

(6) Acceleration Error. The acceleration error at maximum a is given by

$$\theta_a = \frac{\Delta\dot{\omega}}{\omega_n^2} = \frac{56\ (2\pi)}{76^2} = 0.00061 \text{ radians}$$

$$\theta_a = 0.0061 \text{ rad} \times \frac{1 \text{ ft}}{0.003 \text{ rad}} = 0.2 \text{ ft}$$

therefore,

$$\underline{\theta_a = 0.2 \text{ ft maximum}}$$

(7) Velocity Error. The volocity coefficient K_v is:

$$K_v = K_o K_d F(o) = 130 \text{ dB} = 3 \times 10^6$$

where

K_o = VCO sensitivity = 63 rad/sec/volt

K_d = phase detector sensitivity = .5 volt/rad

F (o) = gain of operational amplifier = 100 dB

The phase error due to velocity is:

$$\theta_v = \frac{\Delta\omega}{K_v}$$

The sum of all drifts ($\Delta\omega$) is:

Doppler (max) = ±2.8 Hz × 2π = 118 rad/sec

VCO (5×10^3 ppm) = ±20 Hz × 2π = 126

Total $\Delta\omega$ = 144 rad/sec

then,

$$\theta_v = \frac{144}{3 \times 10^6} = 48 \times 10^{-6} \text{ radians}$$

and

$$\theta_v = \frac{48 \times 10^{-6}}{3 \times 10^{-3}} \sim 0.02 \text{ ft}$$

4. SYSTEM ACQUISITION.

a. Data Link. Using a 205-kHz pcm-fm subcarrier with 256 codes, the interrogator selects any one of 254 transponders (one of the 256 available codes is reserved for internal calibration and 000 is not used). The code length could be expanded to handle any number of transponders.

A bit rate of 2500 bits per second is chosen as a compromise between required transmission time and signal-to-noise ratio. The subcarrier is frequency modulated at an index of 1 in accordance with the bit code, bit timing, and message synchronizing. The subcarrier is phase modulated onto the carrier.

b. Data Link SNR. When the data link subcarrier is on (acquisition), the range data feedback loop is open, thereby decreasing the noise contribution by 5 dB. The carrier-to-noise in the if. at an index of 1 is -8 dB in a noise bandwidth of 1.5 MHz. The data link SNR then is

$$S/N_o = c/N_1 \frac{B_L\ (\mathrm{if})}{B_L\ (\mathrm{DL})} \beta^2 = +17\ \mathrm{dB}$$

where

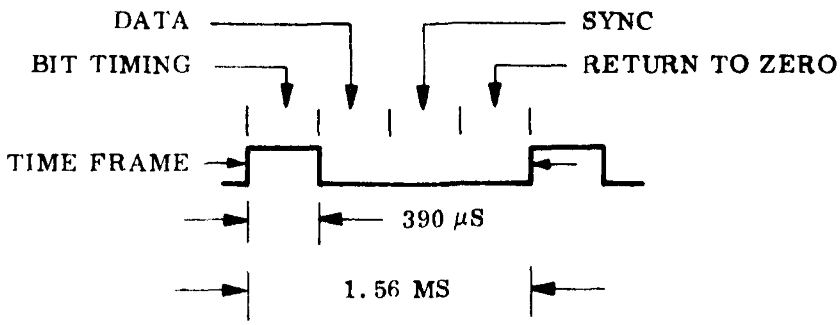
c/N_1 = carrier-to-noise in if. = -8 dB

B (if) = if. bandwidth = 1.5 MHz

B_L (DL) = data link bandwidth = 2.5 kHz

c. Probability of Error. The derivation of the expressions for the probability of error in frequency shift key (FSK) systems is quite involved and will not be given here. Figure 6 plots the carrier-to-noise ratio (dB) versus probability of error.[1] This curve shows that a 12.5 dB SNR in a coherent FSK system provides a probability of error of 10^{-5}.

d. Message Organization and Data Rate. Simplification in hardware and cost can be achieved at the expense of data rate by allocating more time to bit and message synchronization. Data, bit timing, and word sync are organized into a pulse width code with return-to-zero as shown below:



This organization requires a filter bandwidth of $f = \frac{1}{390\ \mu s} = 2.56$ kHz and provides a data rate of $f = \frac{1}{1.56\ \mathrm{ms}} = 640$ bits/sec.

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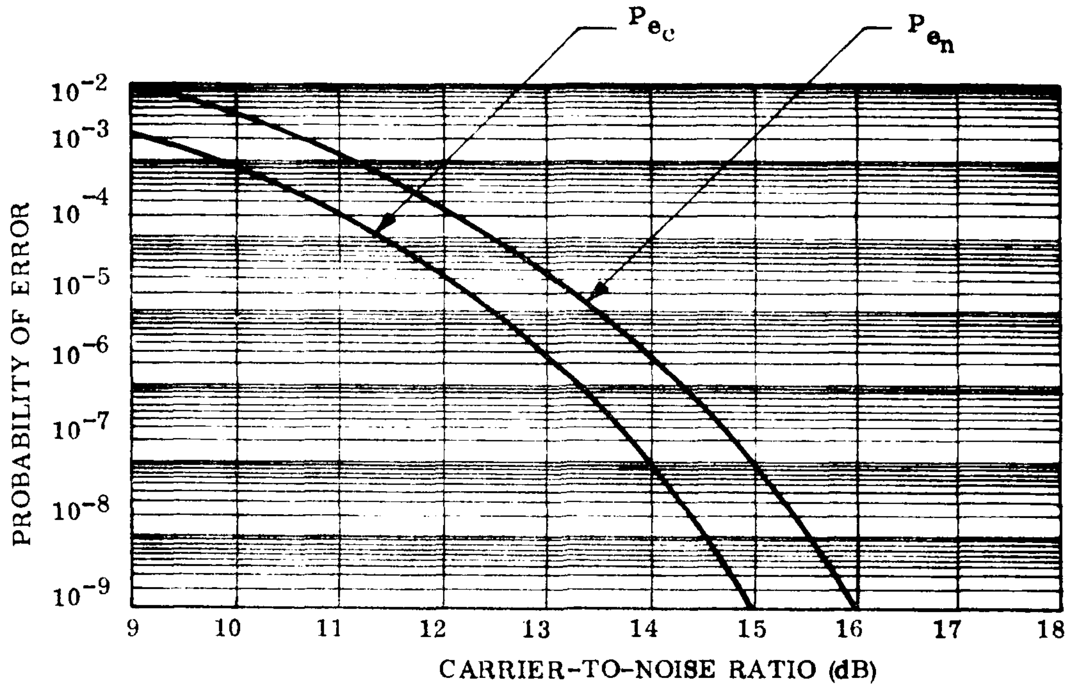
[1] Philip F. Panter, Modulation, Noise, and Spectral Analysis, McGraw-Hill, 1965, p. 714.



Legend:

P_{e_n} = probability of error for noncoherent FSK system.

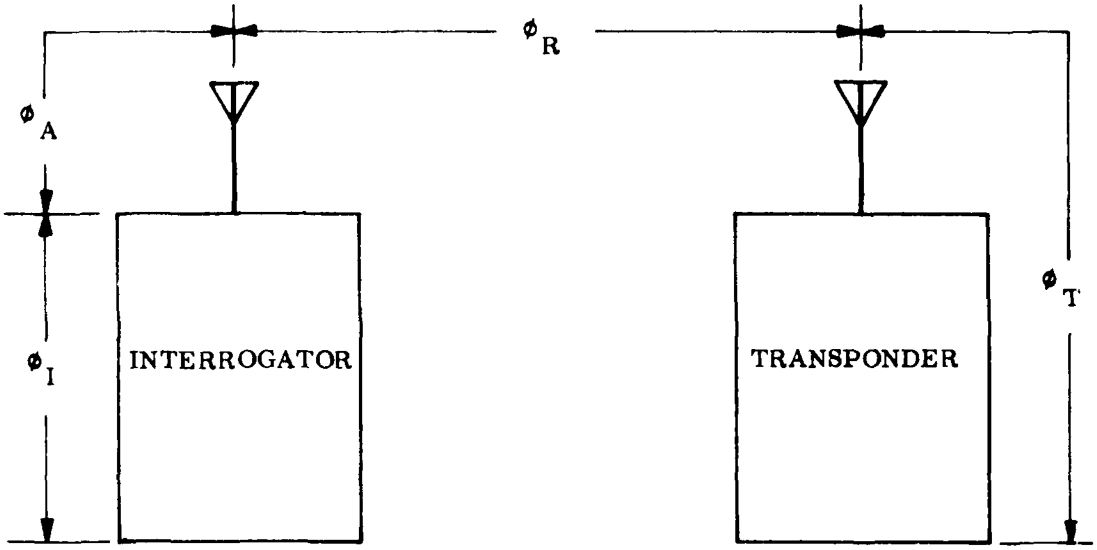
P_{e_c} = probability of error for a coherent FSK system.

Figure 6. Probability of Error in Binary FSK Systems

e. Maximum Code Acquisition Time. The transponder code is 8 bits plus sync for a total of 9 bits. Since the transponder receiver acquisition time is not fixed, it is possible to acquire in such a manner as to lose the first message bit. In this worst-case situation it would require reception of two transponder codes or 18 bits for a total (and maximum) time of 28 rms.

5. ACCURACY CONSIDERATIONS

a. Calibration. Since the RRS determines range based on measurements of signal phase change, fixed phase delays attributable to the interrogator (ϕ_I), to the transponder (ϕ_T), and to the antennas and their cables must be removed from the total measurable phase delay before determining range from the range phase delay (ϕ_R). These fixed equipment phase delays are measured by use of two calibration procedures: an internal calibration, and a loop calibration. Figure 7 provides a functional representation of the phase delays involved.



Where:

$\phi_{measured} = \phi_I + \phi_A + \phi_R + \phi_T$ $\qquad$ $\phi_{measured}$ = total measured phase delay

$\underbrace{\phi_R}_{\text{True Range}} = \underbrace{\phi_{measured}}_{\text{Measured Delay}} - \underbrace{\left[\phi_I + (\phi_A + \phi_T)\right]}_{\text{Calibration Figure}}$ $\qquad$ ϕ_I = Interrogator internal delay

$\left(\phi_A + \phi_T\right)$ = Constant delay of antennas and transponder

Figure 7. Calibration Scheme for Range/Range Rate Subsystem

(1) Calibration Modes. The internal calibration mode measures ϕ_I by phase-locking the interrogator receiver to its own transmitter. In this mode, a signal of the proper frequency is coupled to the interrogator receiver through the antenna circulator. The calibration signal then is processed in the same manner as a ranging signal.

The loop calibration mode measures range between the interrogator and transponder when they are separated by a path of known range. This procedure measures the total combination of phase delays:

$$\phi_{measured} = \phi_R + \phi_I + \phi_A + \phi_T$$

Subtract the known ϕ_R and the ϕ_I as measured during internal calibrate results in:

$$\phi_{measured} - \phi_R\,(\text{known}) - \phi_I\,(\text{int cal}) = \phi_A + \phi_T$$

The transponder is a completely closed-loop, phase-locked device so that its phase delay (ϕ_T) remains stable or constant. Similarly, ϕ_A remains constant as long as the antennas and their cables remain the same electrical length. The combination $\phi_A + \phi_T$ then represents the fixed part of the calibration constant.

The interrogator calibration contribution ϕ_I, on the other hand, may drift slowly with time due to temperature changes or component aging. This drift occurs primarily in those interrogator circuits which are not part of the receiver's feedback loop.

During system operation, interrogator drift is compensated by performing frequent internal calibrations to measure ϕ_I. These values of ϕ_I are compared with the value recorded during loop calibration, and if any change has occurred the computer adjusts the calibration word accordingly.

To summarize, the calibration word is made up of two parts: (a) a constant portion due to phase delay in the transponder, antennas, and cables; and (b) a drift-prone portion due to the interrogator. The constant portion is determined once only, during internal loop calibration. The drift-prone portion is monitored and adjusted frequently by internal calibration. Thus,

$$\left[\text{CAL WORD}\right] = \left[\phi_A + \phi_T\right]_{\substack{\text{loop}\\\text{cal}}} + \left[\phi_I\right]_{\substack{\text{int}\\\text{cal}}}$$

In the operate mode, range phase delay is computed by subtracting the calibration word from the total phase delay measured by the RRS; i.e.,

$$\phi_R = \phi_{\text{measured}} - \left[\text{CAL WORD}\right]$$

b. Calibration Accuracy. If calibration is attempted using a known air link path between interrogator and transponder, the calibration accuracy will be dependent on the air link path ϕ_R during such loop calibration. Since multipath then becomes a great source of error, the system is more accurately calibrated with a cable link of known electrical length. This type calibration need be conducted only once to determine the constant $\phi_A + \phi_T$. It is then set into the computer as a known constant.

This type of loop calibration is a factory or depot-level task, and is not normally attempted in the field. On the other hand, internal calibration can and should be performed several times every hour during system operation in order to compensate for drift in the interrogator. Using the cabling method, the transponders and cables can be calibrated to a maximum error of 1 foot. The interrogator also can be calibrated to this same accuracy of 1 foot.

c. Drift Considerations.

(1) Transponder. The transponder is a completely closed-loop phase-locked device with range modulation feedback of 30 dB. Any drifts would be reduced by that amount. Extensive tests have shown that the transponder drifts are less than the readout capability of 1 foot.

(2) Interrogator. Drifts in the interrogator are removed by the internal calibration scheme. The drifts occur slowly with time due to temperature changes or component aging. With internal calibration performed several times during each mission, a maximum drift value of 1 foot will be assigned to the error budget.

d. Oscillator Stability. The stable reference oscillators used for both the carrier and the range modulation tones are temperature-compensated crystal oscillators requiring no warm-up time or oven power. Frequency stability versus temperature (-40° to +70°C) is $\pm 1 \times 10^{-6}$. Stability versus time is $\pm 1 \times 10^{-8}$/24 hours.

(1) Contribution to Range Error. At 1-ppm oscillator error, the range error at the maximum range of 200 miles is $200 \times 10^{-6} = 1$ ft. With long-term stability of 1×10^{-8}/24 hours, the system could be unattended for 3 months before the oscillator would drift beyond 1 ppm.

(2) Contribution to Range Rate Error. At 1-ppm oscillator error, the maximum rate error occurs at maximum velocity: 5000 ft/sec $\times 10^{-6} = 0.005$ ft/sec.

e. Equipment Noise. The phase error due to equipment noise would be difficult to calculate (and probably inaccurate). Laboratory measurements were taken to detect the presence of noise. For testing purposes the equipment was operated with a signal input at high signal-to-noise ratio ($\approx$ -60 dBm input signal). Any remaining measured jitter was defined as equipment noise. The jitter was observed on an oscilloscope by monitoring the least significant flip-flop of the fine range servo while synchronizing the oscilloscope to the reference. The peak-to-peak time jitter was observed at less than 60 nanoseconds, corresponding to approximately 0.5 ft p-p jitter at a VCO frequency of 7.7 MHz.

The corresponding flip-flop in the range rate servo indicated a p-p time jitter of less than 50 nanoseconds, corresponding to less than 0.01 ft/sec at a VCO frequency of 10 MHz.

f. Multipath Error. A source of range error in any precision CW tracking system arises when the signal is received from more than one propagation path by reflection from the ground or nearby objects. This multipath phenomenon can cause DME phase errors and possible ambiguous range measurement if the phase error is larger than the phase overlap of the modulation frequencies.

Multipath is a function of geometry, the reflection coefficient of the earth's surface and the roughness of the reflection point. The range error or amount of phase shift caused by multipath is also dependent on the modulation index along with the vector addition of the multipath. It is significant to note that with a modulation index of 20 and with the reflection coefficient a maximum of 1, the maximum phase error is $\approx 5°$. This corresponds to a maximum range error of

$$5° \times 5.7 \text{ ft/deg} = 28.5 \text{ ft}$$

This absolute maximum value of α is misleading because it does not indicate typical values. For example, typical values of α measured experimentally as $R = 0.1$ by SHIRAN (AN/USQ-28) tests would cause a phase error of about 0.35°, corresponding to a distance error of

$$0.35° \times 5.7 \text{ ft/deg} = 2 \text{ ft}$$

Some preliminary ground testing of subsystem equipment indicates and supports this number for ground-to-ground measurements.

The maximum error that could ever be observed in range-rate measurements due to multipath would be a change in multipath during the 1-second count period that would cause the phase error on the carrier to change by 90 degrees. This would cause an error maximum of:

$$90° \times \frac{0.02 \text{ ft/sec}}{22.5°} = 0.08 \text{ ft/sec}$$

Carrier multipath normally appears as a slow fading or null in carrier signal strength. The phase error rate of the carrier is therefore quite slow. Not enough data exists to assign an absolute value on this error. If one can assume a reflection coefficient change of 0.1 as typical, then an error of 0.01 ft/sec would be a reasonable number.

g. Atmospheric Refractive Index. The refractive index determines the propagation velocity of electromagnetic waves and also determines the wavelength of electromagnetic waves of a given frequency. In order to correctly convert phase measurement into range, one must know the refractive index. Any error in the refractive index results in a corresponding scale factor error in the range measurement.

The rf atmospheric refractive index varies with pressure, temperature and water vapor content. The accuracy that one can determine these variables will of course affect the ultimate accuracy. Detailed analyses of the problem have been conducted and some practical correction techniques devised. The degree of prediction is about 10 ppm with ground measurements of wet and dry bulb temperature plus airborne altitude measurements. These propagation effects will therefore affect both the range and range rate by about 10 ppm.

One way to minimize this error is the use of multiple transponders spaced g
cally on the range to keep long-range measurements to a minimum. The us
call addressing to the transponders along with the data link verification allo
number of transponders to be used with the interrogator. The low power co
unattended operation capability, portability and low price of the transponder
use of multiple transponders a practical solution to the problem.

h. Range Reference Zero-Crossing Error. The range data is re
crossing of the range reference counter. Since this actual time of reading r
delayed beyond the read signal (R-STOP) by a maximum time t_{max} of

$$t_{max} = \frac{1}{f} = \frac{1}{3.751 \text{ kHz}} = 267 \text{ microseconds}$$

Assuming a maximum aircraft velocity of 5000 ft/sec directly toward or aw
the transponder, the maximum error in range, ΔR_{max}, is

$$\Delta R_{max} = 5000 \text{ ft/sec} \times 267 \times 10^{-6} \text{ sec}$$

Assuming the worst-case in aircraft vector dynamics and a $3\sigma = 267\ \mu s$ prob
determining the time of reading of range, yields

$$1\sigma = 0.45 \text{ ft (max. v)}$$

SECTION III

DESCRIPTION OF RANGE/RANGE RATE SUBSYSTEM

1. CIRIS SYSTEM OPERATION: (See figure 8) The principal elements of the CIRIS integrated reference system include (a) ground transponders appropriately emplaced at precisely known (surveyed) locations and (b) an airborne digital computer properly interfaced with the RRS interrogator and an inertial measurement unit (IMU). Operating under the control of the computer, the interrogator calls up selected transponders and supplies range, range rate and data quality information to the computer. The computer then uses these data to establish the real-time spatial location of the aircraft, updating the IMU as necessary, and thereby maintaining an accurate record of aircraft position and attitude vs. time. The resulting position and attitude data output can be continually displayed in real time, recorded on magnetic tape for post-flight data reduction, or both.

2. RRS SUBSYSTEM OPERATION.

a. Basic Operating Principles. In performing its two main functions, supplying the airborne computer with accurate range rate and slant range data, the RRS makes use of the following basic principles:

(1) Range Rate. The range rate circuitry operates on the principle that the frequency of a cw signal received from a moving source varies in proportion to relative velocity (doppler effect). By maintaining the transponder's carrier frequency at a fixed relationship to the carrier frequency it receives, the doppler frequency is preserved. The transponded carrier frequency received at the interrogator thus can be measured and compared to the carrier as originally transmitted by the interrogator. The difference frequency (two-way doppler) can be converted to a measure of aircraft velocity with respect to the transponder's known location.

(2) Slant Range. The ranging circuits of the RRS operate on the principle that a cw electromagnetic wave propagated through space undergoes a phase change proportional to distance traveled. The phase of selected modulating frequencies transmitted to a transponder and coherently retransmitted back to the interrogator thus can be compared to the phase of the signals as originally transmitted, and the measured phase delay can be converted to a measure of the straight-line distance between aircraft and transponder (i.e., slant range).

b. General Operation. The airborne interrogator interfaces with the onboard computer substantially as shown in the simplified subsystem diagram (figure 9), and communicates with the ground transponders as controlled and commanded by the computer. Operating under computer control, the RRS provides (a) the air-to-ground data link that activates the desired transponder and verifies that the selected

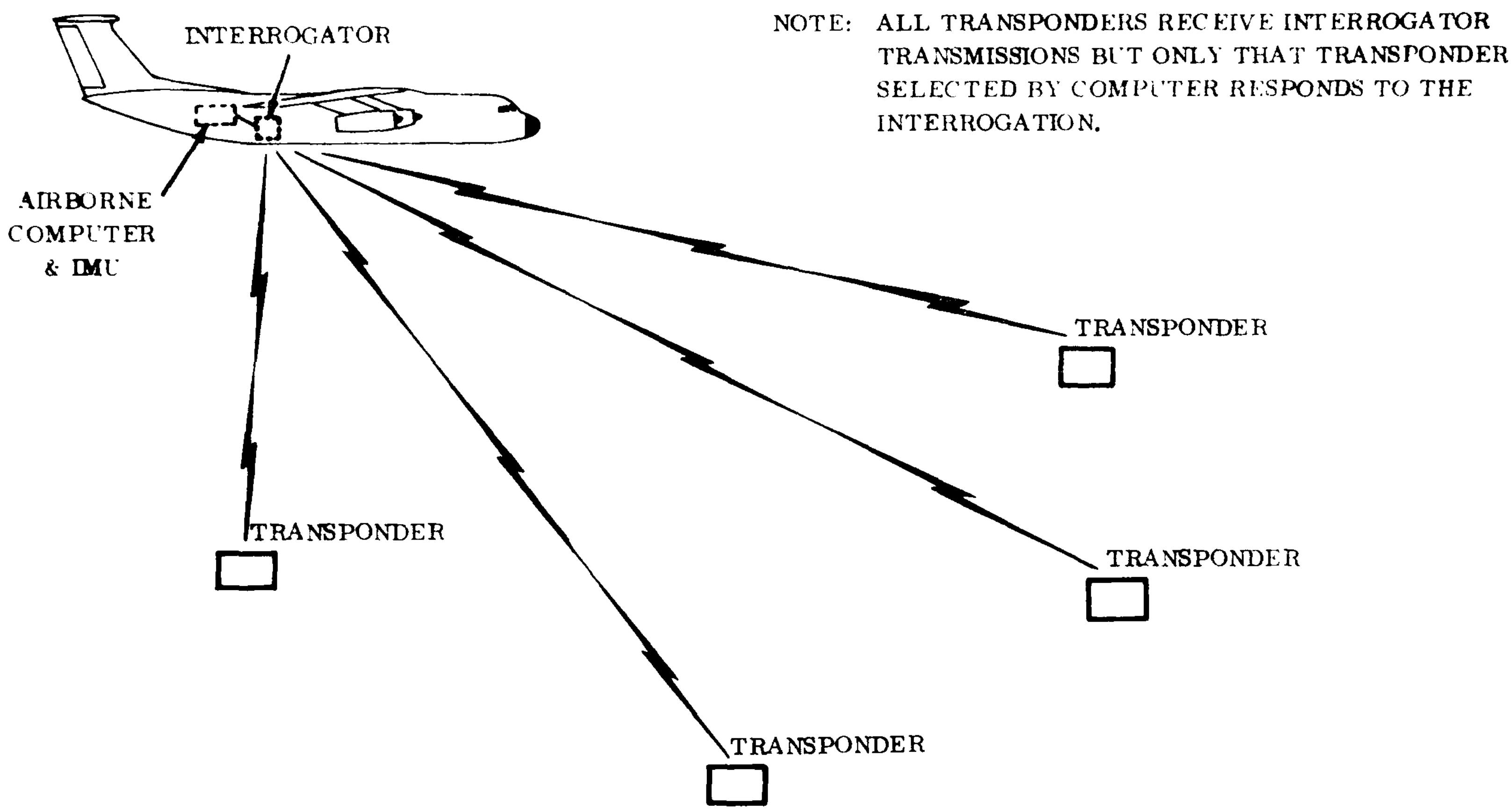


Figure 8. Range/Range-Rate Subsystem (RRS) as Used in CIRIS

transponder responds to the interrogation; (b) a coherent carrier tracking loop for deriving range rate data; (c) a range modulation tracking loop for deriving range data; and (d) data storage registers for holding the range, range rate, and data quality words for sequential readout to the computer on command. Additional RRS features include an in-flight self-calibration capability for the interrogator and two-way voice communications with each ground transponder.

c. Computer-Interrogator Interface. As figure 9 illustrates, the computer-interrogator interface consists of the following twisted-pair control and data lines: (a) power on-off control, (b) four pairs of address lines and associated strobe, (c) two pairs for the transponder-select (ID) data input, (d) one pair for calibrate commands, and (e) 11 pairs for parallel data transfer from RRS to computer. (Section IV contains a detailed description of the computer-interrogator interface.) The general functions of the control and data lines are as follows:

(1) The power on-off control line allows the computer to control the 400-Hz 3-phase primary supply that powers the interrogator. For manual powering, a manual ON/OFF switch on the interrogator by-passes this line.

(2) The four address lines provide up to 16 4-bit BCD commands for controlling the interrogator. For example, binary 1111 (address 15) is used as the FIX command that initiates every transponder interrogation; addresses 1 through 9 address various parts of the interrogator's data storage register; address 10 starts the range rate measurement, and address 11 stops it.

(3) The strobe is the computer's "device command encode" pulse which (a) strobes the 4-bit addresses into the interrogator's decode matrix, (b) strobes the transponder ID into the interrogator's encode/decode register, (c) strobes the calibrate command into the interrogator, and (d) strobes the data onto the 11 output lines for entry into the computer.

(4) The transponder ID input lines are used to enter the 9-bit transponder ID (select call for transponder to be interrogated) into the interrogator. One line is used to enable a gate through which the ID bits (8-bit binary number plus even parity bit) are serially strobed.

(5) To initiate an interrogator calibration sequence, the computer makes the calibrate line a logic "1". The strobe line then executes the calibrate command.

(6) The 11 data output lines permit the computer to strobe range, range rate and data quality information out of interrogator storage and into the computer in 11-bit segments. The range data acquired in each interrogation consists of four 11-bit words, the range rate data consists of two 20-bit words (each shifted out

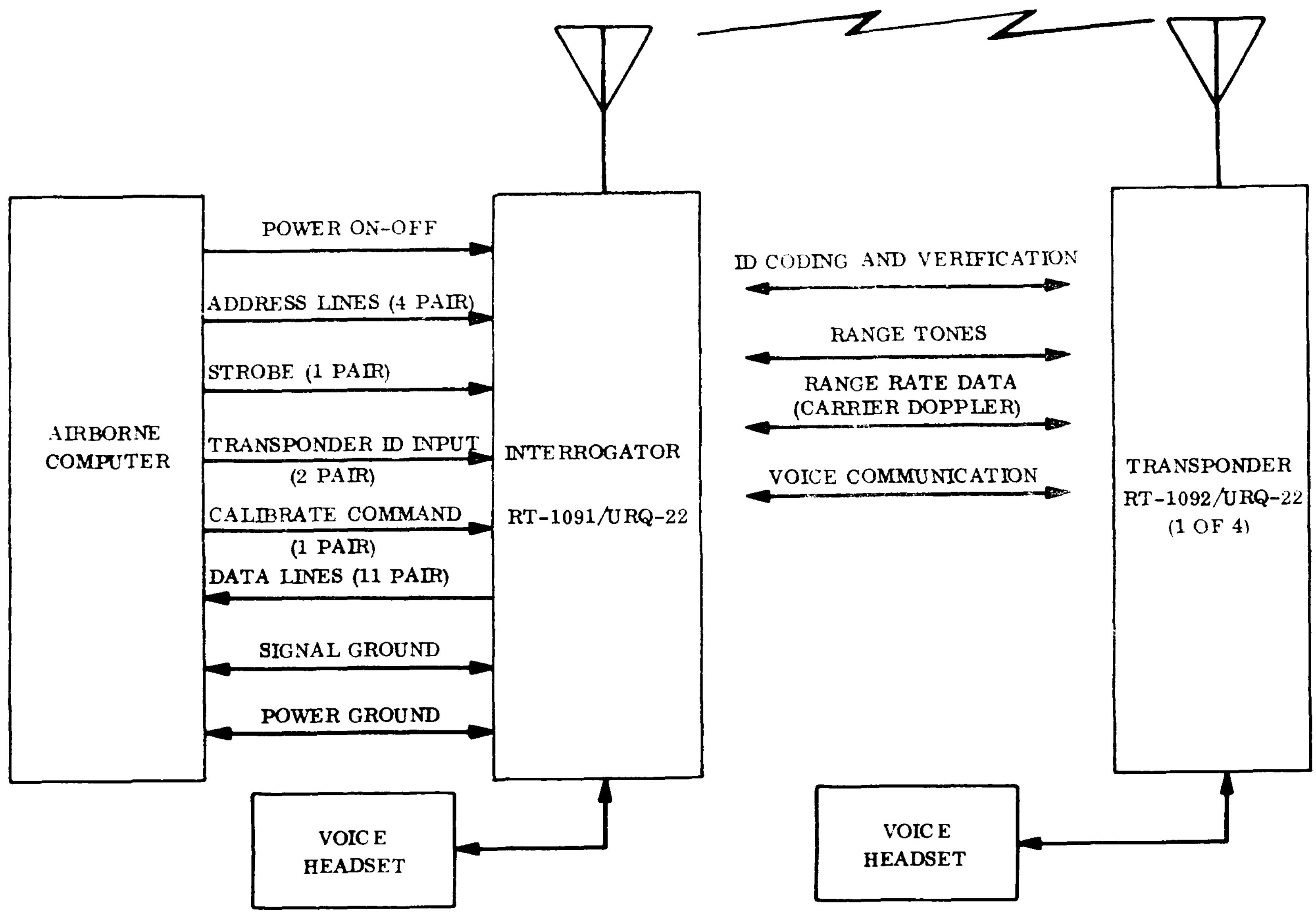


Figure 9. Simplified Block Diagram of Range/Range Rate Subsystem

in two increments), and the data quality information consists of 7 bits, each of which checks an essential operation in the interrogation sequence.

d. Sequence of Events in Typical Interrogation. Figure 10 shows the approximate times required for various parts of a typical interrogation. For the worst case, for example, 1008 ms would be needed. The sequence of events in a typical interroation is descrived below. (See figure 11.)

(1) Using the ID input and gate lines and the strobe, the computer shifts the select call ID of the transponder to be interrogated into the interrogator's encode-decode logic; then, using the address lines and the strobe, it issues the FIX command (address No. 15) to initiate the interrogation sequence. To prevent entry of false data or noise, (a) strobe data enters the transponder ID line only when its gate is enabled, and (b) while ID bits are being strobed in, the unused "0000" address is applied to the address lines.

(2) The encoder logic in the interrogator checks the transponder ID word for correct (even) parity; if satisfactory, it turns on the interrogator transmitter and activates the data link subcarrier circuits. (If parity does not check, the interrogator notifies the computer by placing a parity error bit in the data quality word register.)

(3) When activated, the interrogator data link circuits convert the 8-bit transponder ID code word into a pulse-width code and add a sync pulse. The code is generated at a 5-kHz clock rate where (a) each data bit is four clock times wide; (b) the first clock time is used as a timing bit for the acquisition and phase-lock loop, and is always high; (c) the second clock time is used for the data bits (high = "1", low = "0"); (d) the third clock time is used for the sync bit, where high = sync; and (e) the fourth clock time is the return-to-zero period that is always low. (See figure 12) This pulse-width code frequency-shift-keys (FSK) a 205-kHz subcarrier oscillator, and the subcarrier in turn phase-modulates the interrogator carrier. The eight data bits (ID code word) plus sync are repeatedly transmitted over the data link until verification is obtained from the selected transponder.

(4) Within 25 ms, all transponders within line-of-sight range of the interrogator acquire and phase-lock to the interrogator carrier; within the next 28 ms they lock the subcarrier demodulator to the incoming 205-kHz subcarrier oscillator (SCO) and also sync to the data link bit-rate frequency. All transponders decode the 8-bit select-call word and check it for code match.

(5) The transponder whose ID matches the word transmitted responds by (a) turning on its transmitter, (b) reencoding its own 8-bit ID into pulse width modulation form, and (c) using the resulting signal to frequency-modulate a 130-kHz subcarrier oscillator which in turn phase-modulates the transponder carrier. The FSK subcarrier is transmitted for a fixed time period of about 35 milliseconds.

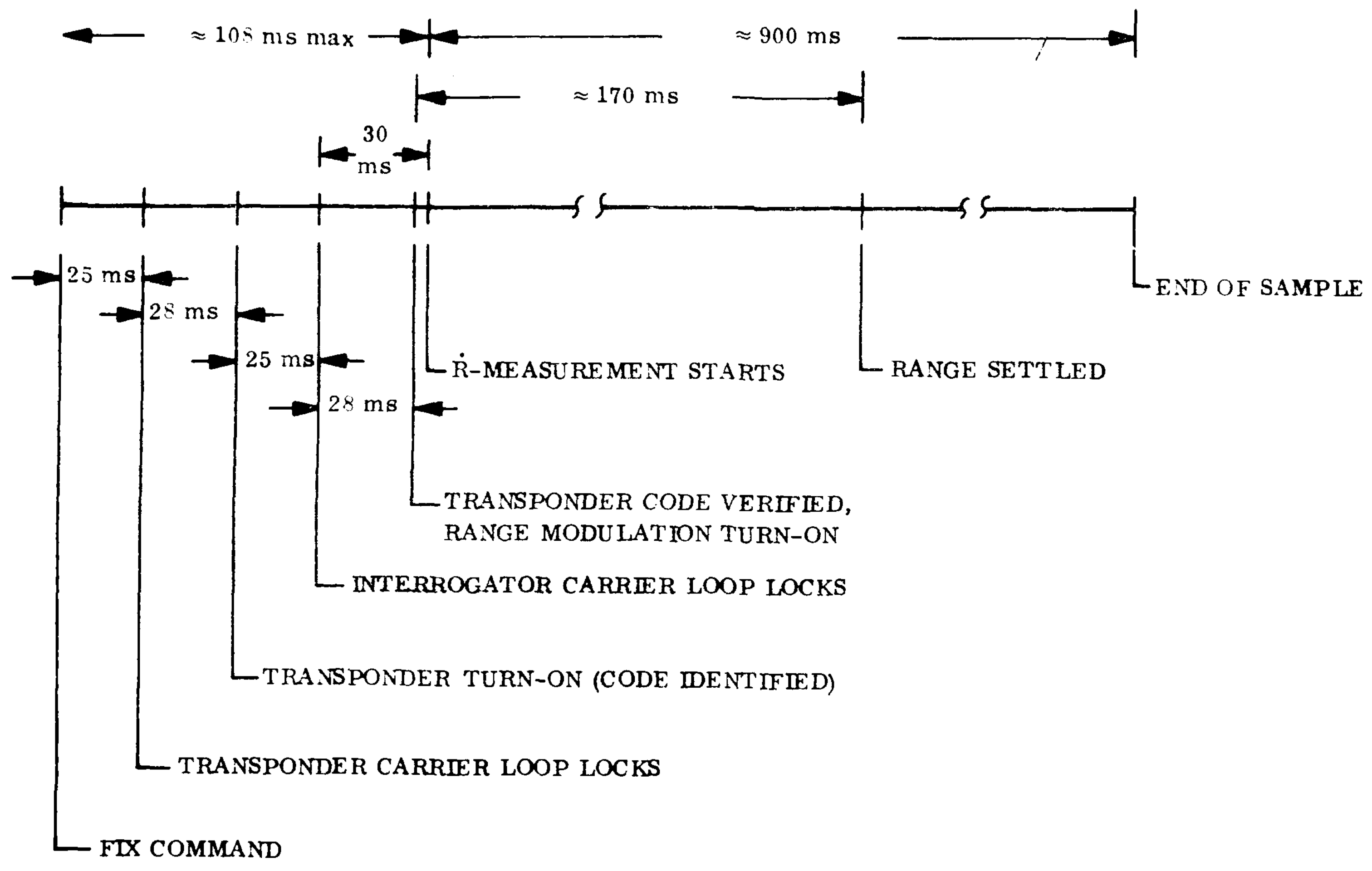


Figure 10. Worst-Case Timing for Sequence of Events, Typical Interrogation

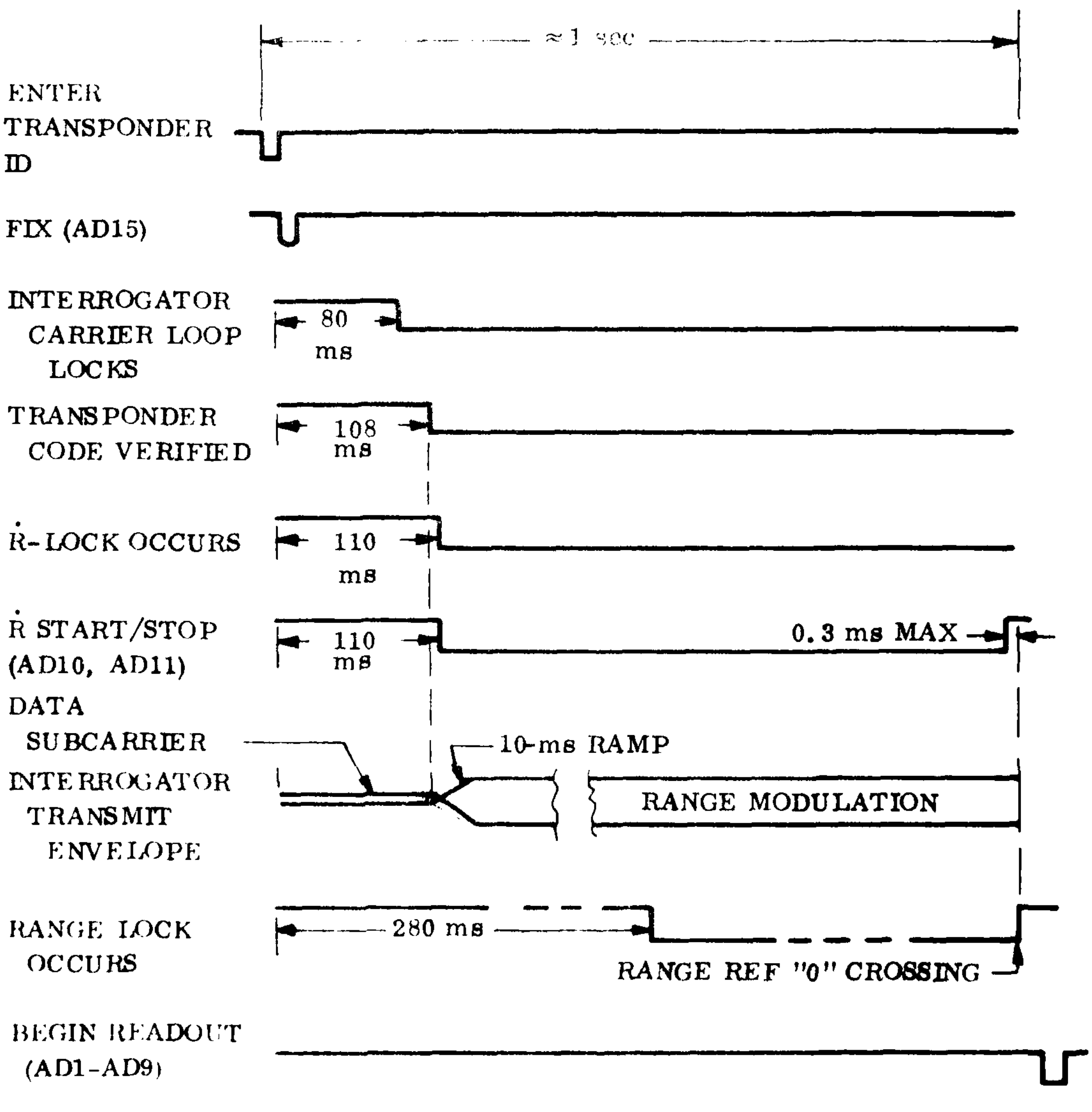


Figure 11. Interrogation Sequence Timing Relationships

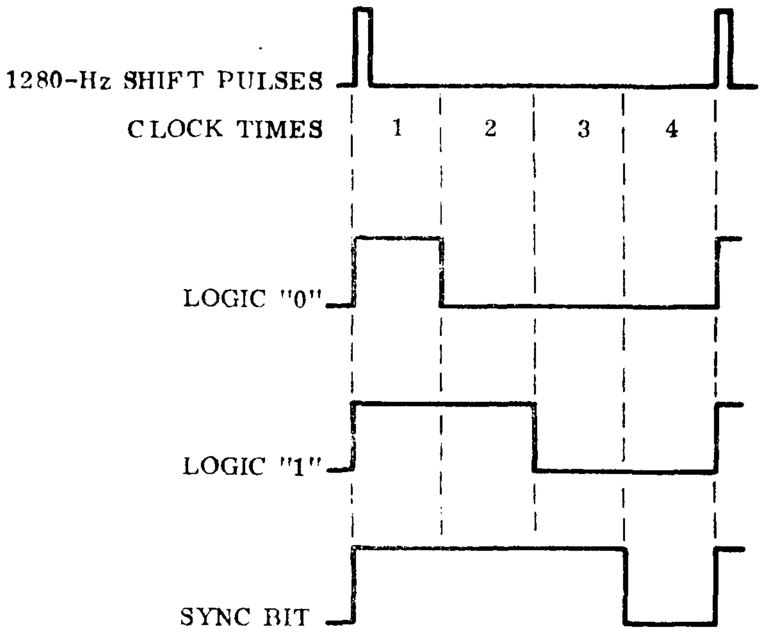


Figure 12. Pulse Width Coding Waveforms for Transponder ID Word

(6) Within 25 ms after transponder transmitter turn-on, the interrogator receiver acquires and phase-locks to the transponder carrier; carrier loop lock, in turn, activates the range-rate ($\dot{R}$) servo loop and its associated 30-ms delay circuit and deactivates the 205-kHz subcarrier oscillator. Following the 30-ms delay period (which allows the $\dot{R}$ servo loop to settle) the interrogator notifies the computer by inserting an $\dot{R}$ LOCK bit in the data quality register. The computer (which is continually sampling the data quality register using readout address No. 9) initiates a range-rate count by activating address No. 10 ($\dot{R}$ START). Also at this time, the interrogator logic sets an $\dot{R}$ LOSS flip-flop which will be toggled if loop lock is lost at any time during the range rate count. The state of this flip-flop forms part of the data quality word, and is examined at the end of the interrogation cycle to assure that the range rate data is valid.

(7) While the $\dot{R}$ servo loop is settling, the transponder ID verification circuits in the interrogator are demodulating the FSK subcarrier, the FSK bit-rate loop is being synchronized, and the transponder ID being received is decoded and compared for match with that which was transmitted. When the FSK loop locks, an FSK LOCK bit is set into the data quality register. When code match is verified, a verification bit is set into the data quality register; when $\dot{R}$ servo loop has settled, the range modulation circuits are activated and the data link circuits in the interrogator are deactivated.

(8) Activation of the range modulation circuits causes the range modulation baseband (a composite of four ranging tones) to be "ramped" or gradually impressed on the interrogator carrier. The modulation index is linearly increased over a 10-ms period, thereby preventing the loss of carrier loop lock that would occur if the modulation were to be abruptly applied at full index.

(9) In the transmitting transponder the end of the 35-ms FSK transmit time period deactivates the 130-kHz FSK subcarrier and closes the data feedback gate. The range modulation now contained in the received carrier is demodulated, filtered, recombined, and remodulated onto the transponder carrier.

(10) The interrogator receiver demodulates the range modulation now contained in the transponder transmitter signal, then filters out the individual range tones and supplies them to the range translator and servo circuits. When all four range data servos lock up, a delay circuit is activated to allow the range measurements to settle; if any range servo drops out of lock during the delay period, the delay is repeated, thus preventing any range measurement until all servos have had time to settle. When all range servos remain locked throughout the delay period the interrogator inserts an R LOCK bit into the data quality register.

(11) The computer terminates the range-rate measurement period by issuing the $\dot{R}$ STOP command (address No. 11). The next zero crossing of the range reference then terminates the interrogation by turning off the interrogator transmitter. As soon as the transponder loses carrier loop lock it turns off its transmitter also.

(12) The four range partials (four 11-bit words), the range rate data and range rate reference (two 20-bit words), and the data quality bits are immediately available for readout by the computer. These data words are presented to the 11 data lines in response to readout addresses 1 through 9. If the interrogation sequence was satisfactory in all respects, the data quality bits will consist of (a) a "1" bit in the FSK LOCK location, indicating that FSK lock was achieved during the interrogation; (b) a "0" bit in the PARITY ERROR location; (c) a "1" bit in the verification location; (d) a "0" bit in the R LOSS location; (e) a "1" bit in the R LOCK location, (f) a "0" bit in the "malfunction" location, indicating that the monitored operating indicators (power supply voltages, etc.) were all within tolerance; and (g) a "1" bit in the $\dot{R}$ LOCK location.

The computer program, interrogator circuits, and transponder circuits provide a number of corrective responses when certain features of an interrogation sequence are unsatisfactory. If a parity error is noted in the transponder ID, for example, the computer may try several times to shift in the ID code word and issue a FIX command; if parity is still in error, it informs itself that a malfunction exists. Also, after a transponder has responded to carrier acquisition, FSK lock, and ID code match by turning on its transmitter, any carrier dropout (loss of lock) greater than 10 ms causes the transponder to revert to the standby mode. Similarly, loss of lock of 10 ms or more in the carrier tracking loop of the interrogator receiver re-initiates the interrogation

of the selected transponder; however, if R LOCK and verification are not obtained within a reasonable time (approximately 300 ms), the computer program can detect and store this fact, and normally will proceed to the next transponder to be interrogated.

e. Transponder ID SCO and Verification Loop. The digital data link circuits consist of (a) ID encoder and FSK subcarrier oscillator (SCO) circuitry in the interrogator, (b) data link acquisition circuitry in the transponder, (c) verification encoder and FSK SCO circuitry in the transponder, and (d) data acquisition and verification decoder circuitry in the interrogator. These circuits are activated at the beginning of each interrogation, then are shut down when the transponder transmitter is turned on and its ID is verified by the interrogator.

Each fix command is immediately preceded by receipt of the data word representing the transponder the computer has selected for interrogation. The ID word is checked for correct parity. If parity checks, the FIX command that follows is permitted to turn on the interrogator transmitter. When the FIX command becomes effective, the 8-bit ID register in the interrogator encoder section is transferred to the control of the bit rate oscillator and associated programmer-encoder logic. The programmer consists of a divide-by-4 circuit and coincidence gating which, when driven by the clock signal, produces the shift clock pulses and the timing bits for use in serially encoding the 8-bit ID and adding the sync bit. The encoder logic receives each ID bit as it is shifted out of (and recirculated back into) storage and, using the timing pulses supplied by the programmer, converts them into a pulse-width code as illustrated in figure 12.

The resulting pulse train shifts the frequency of the data link SCO accordingly, and the subcarrier in turn in transmitted to all transponders. In the transponder decoder circuits, the subcarrier is demodulated and (a) the leading edge of each data bit transmission, which recurs regularly at a 1280-Hz rate, is used to synchronize the timing counter; (b) logic "1"s are identified by strobing the second clock time, where if the level is high the data bit is either a "1" or a sync; and (c) sync pulses are identified by strobing during the third clock time.

As the logic "1"s and "0"s are decoded they are shifted into an 8-bit register. The contents of the register are compared with the "hard-wired" ID code assigned to the transponder. If the two words match bit-for-bit and the received word is preceded and followed by a logic "1", the ID word is recirculated and gated into the transponder's encoder logic section. The transponder's transmitter then turns on, and the transponder encodes and continually transmits its ID for 35 ms. The interrogator receiver acquires, locks to, and decodes the transponder's data link subcarrier using the same decode methods. As the received ID code is shifted into an 8-bit register in the interrogator decoder logic, the bits are continuously sampled for comparison with the ID originally shifted in by the computer. If at any time the two words match bit-for-bit, verification (code match) is achieved, and the entire data link is disabled (having served its purpose). The range rate and range measurement sequences then are begun.

f. Range Rate (Carrier Tracking) Loop. A stable range rate ($\dot{R}$) reference oscillator in the interrogator generates a basic frequency that is multiplied to obtain the transmitter carrier frequency. This oscillator also generates the range rate reference. When the computer issues the transponder ID and FIX command, the interrogator transmits on its assigned carrier with the data subcarrier as its modulation.

At this time, the selected transponder is operating in a receive-only mode, using a first local oscillator (LO) signal derived from a voltage-controlled oscillator (VCO). The transponder transmit signal is derived by amplifying a portion of the receiver first local oscillator to a nominal 4 watts. When an interrogator transmission having the proper ID is received, the transponder turns its transmitter on; (i.e., it switches on the supply voltage to the power amplifiers and closes a diode switch).

Prior to acquisition, the first LO frequency is being varied ±40 kHz at a 40-Hz rate by applying a triangular sweep signal to the VCO control. (This feature permits rapid acquisition of the interrogator carrier while maintaining a narrow receiver bandwidth.) As soon as the interrogator carrier is acquired, the lock detector used in the receiver disables the VCO sweep circuit; the VCO then is controlled by the carrier tracking error signal. Thus, the transponder acquires and phase-locks to the received carrier. Since the transmitter derives its carrier signal from the first LO, the received carrier is tracked at a phase-coherent fixed ratio, therefore any received carrier shift due to doppler will be reproduced proportionally in the transmitted carrier at the same ratio.

The interrogator receiver acquires and phase-locks to the transponder carrier in substantially the same manner, with the first LO signal supplied by a VCO controlled by a phase detector, forming a phase-locked loop. The VCO frequency thus is coherently related to the original reference carrier transmitted by the interrogator; however, it is shifted in frequency by the magnitude of the round-trip doppler shift resulting from the velocity of the interrogator relative to the transponder. The VCO frequency may therefore be appropriately translated for use in driving the range rate data counter via the range rate data translator.

The FIX command starts an interrogation sequence and also resets the range rate data and reference counters. The computer then allows 100 milliseconds for carrier loop lockup, at which point the $\dot{R}$ START address line goes true, simultaneously enabling input gates for both counters. The counters then receive and hold the doppler and reference counts (two 20-bit words) for transfer to the computer upon demand after the interrogation is completed. Additional logic detects range rate servo loop lockup ($\dot{R}$ LOCK signal) and any loss of lock during the range rate data count ($\dot{R}$ LOSS signal).

g. Range (Modulation Tracking) Loop. A frequency synthesizer in the interrogator generates the range modulation that is transmitted to the transponder, retransmitted by the transponder, then received and demodulated by the interrogator receiver so that the round-trip phase delay can be measured and converted to range.

The range measurement feature of the RRS is based upon the principle that sine wave modulation on an rf carrier propagated through free space undergoes a phase shift directly proportional to distance traveled. This phase shift is independent of carrier frequency, and can be measured by means of an electronic phasemeter and converted to line-of-sight distance (slant range).

For nonambiguous measurements, the modulation wavelength must be long enough so that the largest measurable phase delay (i.e., one-half wavelength) corresponds to a large range increment. For good resolution, however, a short wavelength is needed so the smallest measurable phase delay corresponds to a small increment of range. In the RRS, four harmonically related frequencies (tones) are used, wherein the highest ranging tone establishes the range resolution, the lowest establishes the maximum range increment of the system, and the two in-between tones permit the ambiguities to be successively resolved. The characteristics of the four ranging tones are:

Designation	Frequency (kHz)	Half Wavelength (ft)	Resolution (ft)
Fine (FN)	240.053	2,048	1
Intermediate (INT)	30.007	16,384	8
Coarse (CS)	3.751	131,072	64
Very Coarse (VC)	0.469	1,048,576	512

Because of the filtering problems that would be involved if frequencies of such wide variation were to be transmitted and coherently tracked, the three lower tones are folded around the fine tone using digital mixing techniques, producing the following range modulation frequencies:

Symbol	Frequency (kHz)	Deviation	Modulation Index
D1'	240.053	FN	20
D2'	270.059	FN + INT	2
D3'	243.804	FN + CS	2
D4'	235.833	FN - (CS + VC) = FN - CS - VC	2

After folding, the four frequencies listed above are linearly added to produce the composite range modulation baseband. Because of the large modulation index used for the fine tone (which minimizes multipath error), the modulating signal is applied gradually over a 10-ms "ramping" period to prevent loss of carrier loop lock in the receivers.

All transponders that acquire and phase-lock to the interrogator carrier demodulate and filter out the modulation frequencies, recombine them, and apply the resulting baseband signals to their respective phase modulators. However, only the transponder whose transmitter was turned on by the ID word actually retransmits the range modulation, and since the transmitted signal is used also as the first LO for the transponder

receiver, a "modulation tracking" effect is achieved within the receiver. (During shop alignment, the phase of each modulating frequency is adjusted to provide negative feedback within the receiver circuits of the transponder.)

In similar fashion the interrogator receiver demodulates and separates the four ranging frequencies, recombines them, and uses the resultant signal to phase-modulate the VCO signal that generates the receiver first LO, again providing a negative feedback type modulation tracking loop. The individual tones derived from the demodulated transponder signal (which are now designated D1 through D4, corresponding to the D1' through D4' signals originally transmitted) are also translated to a common range servo frequency through use of digital mixers and phase-lock loops. The D1' through D4' (originally transmitted) frequencies are used as references. The range servos themselves are phase-lock loops incorporating 11-bit counters in their feedback loops. Each range servo thus divides the phase delay it measures into 2048 parts; further, after a suitable settling time each servo is stopped as its reference reaches a zero crossing, automatically subtracting the reference so that the contents of its feedback counter is a digital representation of the phase delay for that range tone. The zero crossing of the reference also shuts down the interrogator transmitter, terminating the interrogation. Loss of carrier loop lock in the transponder receiver then causes the transponder to turn off its transmitter and revert to its standby (receive only) state.

Each of the four range servos presents the contents of its range data counter to the 11 data lines in the sequence called for by the computer. The computer program processes the 11-bit "range partials" from each servo using a subroutine that employs (a) each range partial in turn, (b) the current calibration data for the interrogator, (c) the stored calibration data for the transponder interrogated, and (d) index of refraction correction data. The computation subroutine includes an ambiguity resolving algorithm which, with the corrections indicated above, finally produces a corrected composite range word representing the slant range to the transponder interrogated.

h. Voice Communication Link. The voice communication link provides direct radio communications between the interrogator and the transponders. The transponders simultaneously receive the interrogator transmissions, but they can communicate directly only with the interrogator, not with one another. Further, voice communication circuits are locked out automatically during ranging operations.

3. FUNCTIONAL OPERATION OF INTERROGATOR. The interrogator operates with an airborne computer while performing its primary task of providing a radio link to one or more transponders. The principal circuit sections of the interrogator consist of the following: (See figure 13)

- The input command and data control logic, which responds to control signals issued by the computer and its associated control/display panel.

- The transmitter section, which generates the transmitter carrier plying and amplifying the reference frequency f_0.

- The receiver section, which acquires and phase-locks to the trans carrier and demodulates the transponder transmissions.

- The frequency-shift-keyed subcarrier oscillator (FSK SCO) sectio (a) generates the FSK subcarrier for transmission to the transpon interrogated, and (b) filters and decodes the FSK subcarrier conta the transponder transmissions.

- The range rate ($\dot{R}$) data handling section, which uses multiplicatio divisions of the reference frequency f_0 and the output of the VCLO the received carrier to measure range rate.

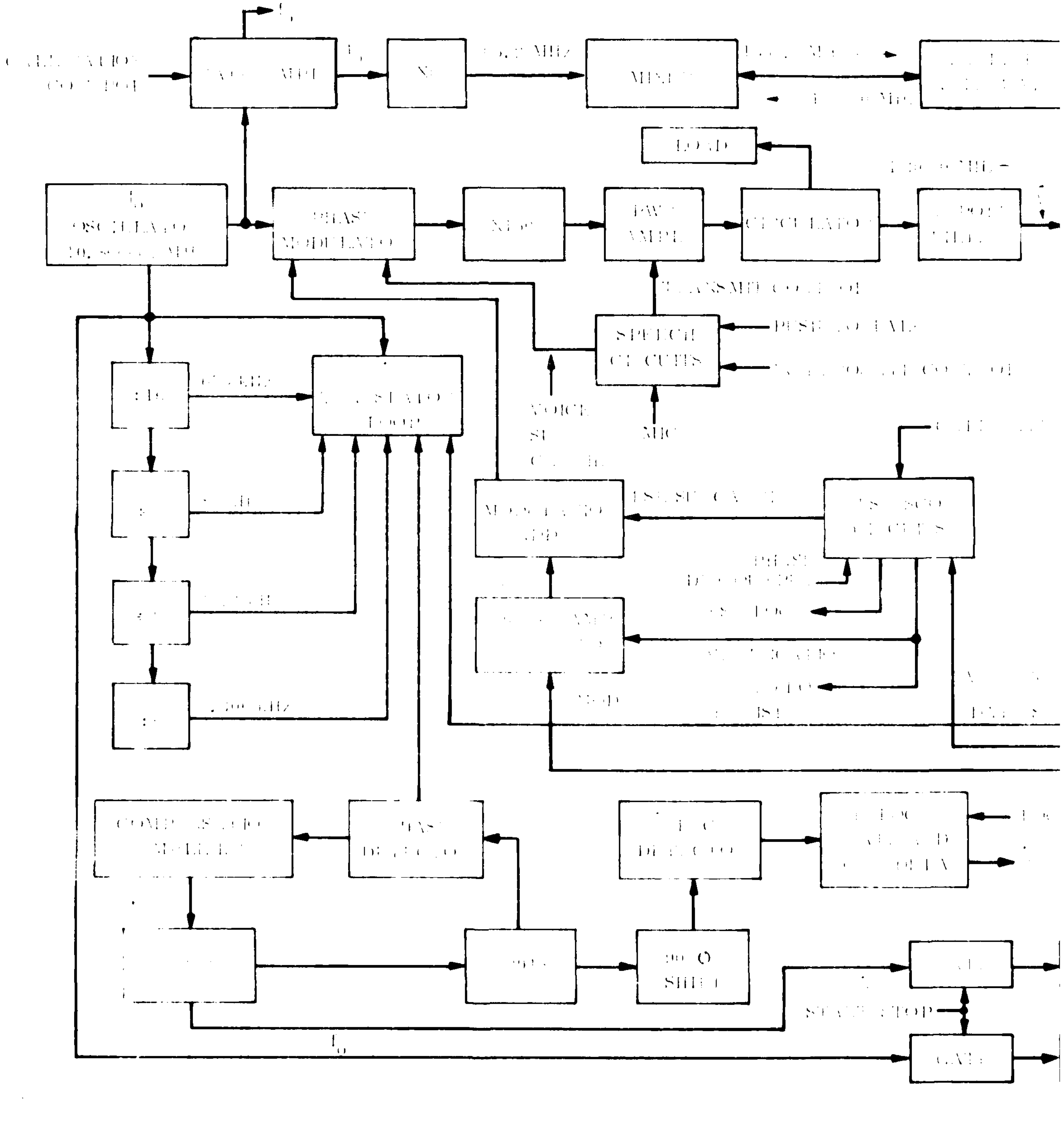
- The range data section, which produces the range modulation base transmission and compares the range references with the received to produce slant range data.

- The calibration section, which (a) produces a simulated received (b) activates a simulated received FSK subcarrier circuit, and (c) a simulated interrogation, thereby providing a means for measuri internal phase delay of the interrogator circuits.

- The voice communication circuits.

a. Control and Interface Logic. The control and interface logic r to the computer commands and provides the necessary control functions for ID word followed immediately by the FIX command (address AD15). If the ID checks satisfactorily, the FIX command is issued, appropriately resetting va logic circuits and setting an interrogator control flip-flop which turns the tra on. (The interrogate control flip-flop is reset at the end of the interrogation zero crossing of the servo reference, thus turning the transmitter off.) Dur interrogation, $\dot{R}$ START (AD10) and $\dot{R}$ STOP (AD11) control the duration of th rate measurement. Following each interrogation the range, range rate, and quality words are presented to the computer in response to addresses AD1 th AD9.

When the computer issues a calibrate command the calibration control circui activated, causing the interrogator to interrogate itself. All interrogate fun are performed during a calibration sequence, yielding a separate phase delay surement for each range modulation frequency and a range rate measuremen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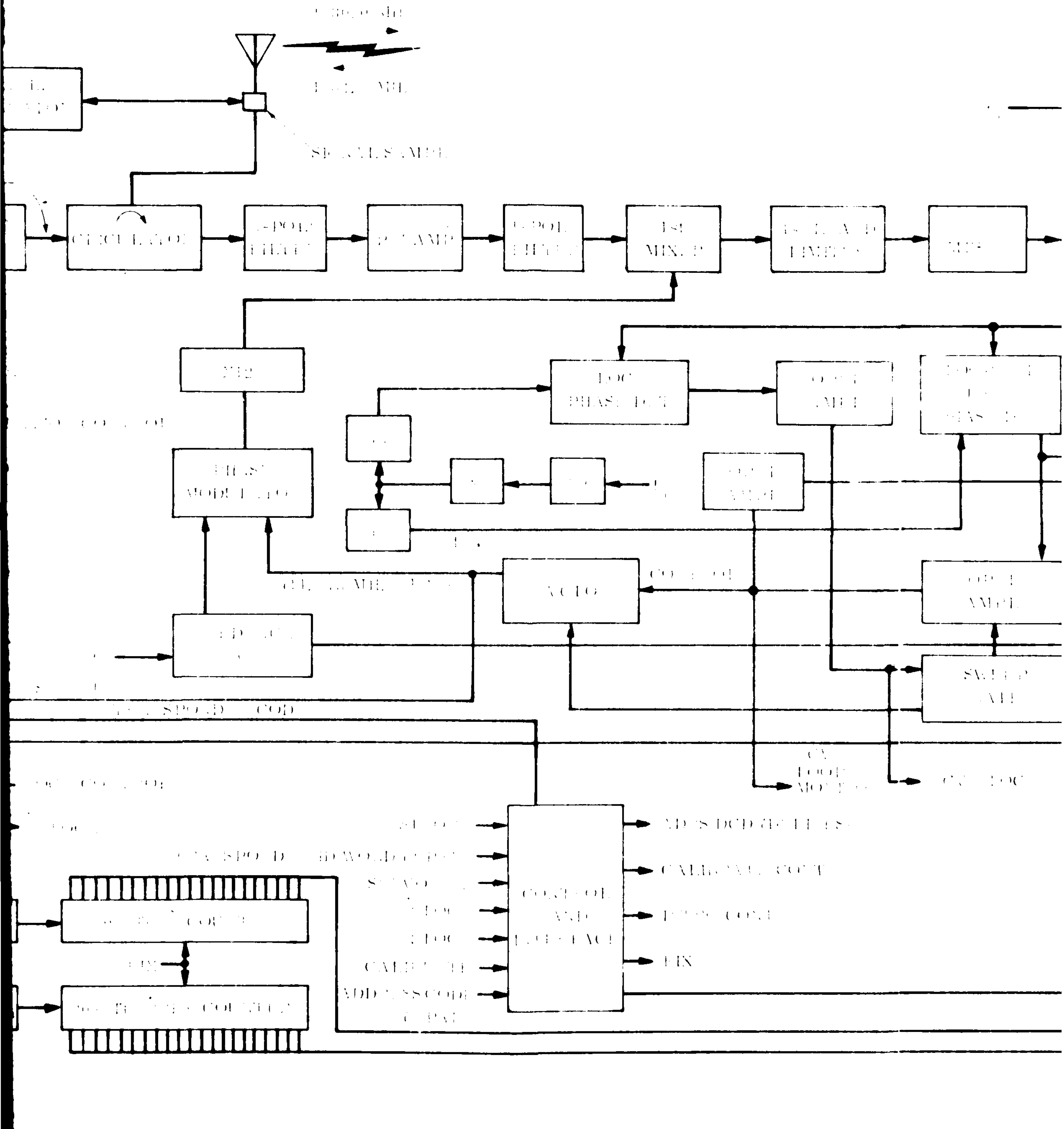
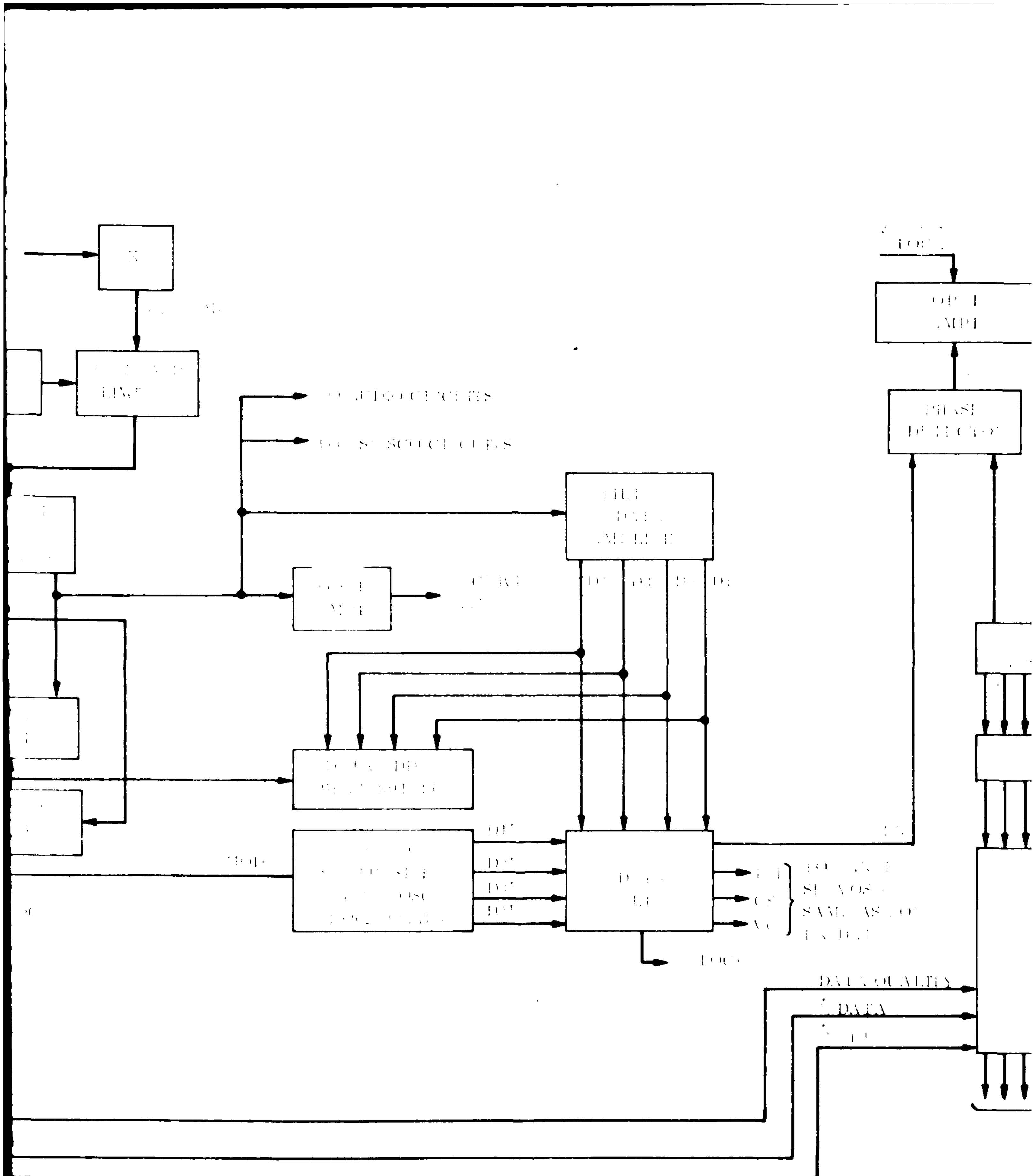


1st
SWEEP
AND



Figure

Diagram

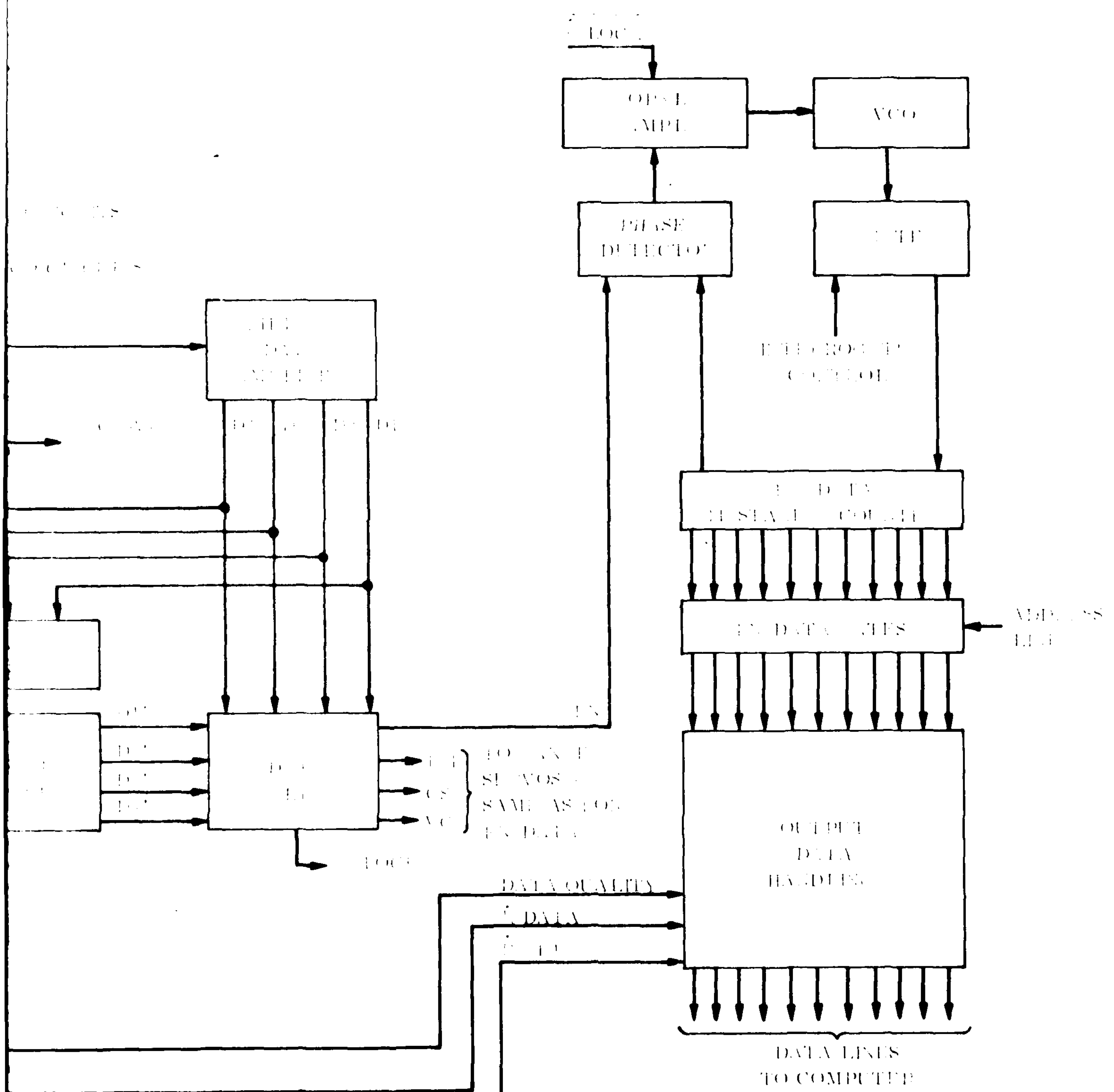


Figure [illegible]5. Functional Block Diagram of Interrogator

The computer reads out the calibration data in the same manner as for a transponder interrogation, uses the data quality and range rate readouts as a confidence check, and stores the range calibration words for use in correcting each range measurement.

b. Transmitter. The transmitter section consists of the following elements:

(1) A stable, temperature-compensated, crystal-controlled oscillator that generates the fundamental frequency (f_o) for the system.

(2) A phase modulator that varies the phase of the f_o oscillator output in accordance with a modulating signal (FSK SCO, range modulation baseband, or voice SCO, as applicable).

(3) The frequency multiplier, amplifier, and filter circuits required to generate the carrier frequency (1630.0 MHz).

Both the fundamental frequency and initial modulation index are multiplied by 150 for transmission to a transponder. The final amplifiers are controlled by the interrogate control flip-flop (interrogate mode) or by the push-to-talk switch (voice communications mode). The f_o oscillator serves also as the master oscillator for the range rate measurement circuits. For this function the f_o signal is supplied directly to the range rate reference counter, and is sequentially divided for use by the range rate data translator. The f_o signal is used again to derive the second LO frequency and correlation detector reference, both employed in the receiver circuits.

c. Receiver. The receiver section contains a dual-conversion radio receiver employing a coherent-carrier phase-lock loop and a correlated lock detector. The output of the loop phase detector is low-pass filtered to produce the control for the loop VCLO (voltage-controlled local oscillator) whose output is multiplied by 128 to produce the first LO frequency. The output of this phase detector also contains any signals phase-modulated onto the received carrier; i.e., the FSK subcarrier prior to transponder verification, the range modulation after verification, or the speech subcarrier during voice communications.

The nominal 1564.8-MHz carrier signal received from a transponder passes through the isolating circulator (A1A7), is filtered and preamplified, then is mixed with the first LO signal (nominally 1494.0 MHz). Prior to acquisition and carrier loop lock, the VCLO is swept by a 40-Hz triangular control signal, thereby varying the first LO frequency ±40 kHz. As the LO frequency crosses the point where the difference frequency between the second if. and reference is within the loop capture range, the correlated lock detector produces a lock control output that disables the VCLO sweep circuit and enables (unsquelches) the range rate counter phase-lock loop.

The first if. signal (nominally 70.633 MHz) is amplified and mixed with the 54.333-MHz second LO frequency, yielding a 16.3-MHz second intermediate frequency. This second if. signal is amplified and limited, then phase-compared with a reference (16.3 MHz) obtained by first halving, then tripling the basic f_o frequency. The reference signal is shifted −45 degrees for use with the loop phase detector, and is shifted +45 degrees for use with the loop lock detector.

As the carrier loop acquires and phase-locks to the received carrier the averaged loop detector output adjusts the frequency of the VCLO so that the second if. signal is maintained in quadrature with the 16.3-MHz reference. The first LO frequency then "tracks" (follows any variation in) the received carrier, thus keeping the intermediate frequencies centered within the passbands of the if. amplifiers. Also, when phase-locked the VCLO output varies from its center frequency by an amount proportional (1:128) to the two-way doppler frequency; i.e., the transponder is similarly phase-locked to the interrogator carrier and is coherently retransmitting. Further, when the carrier loop locks, the lock detector output is maximum (signals being phase-compared are in phase).

The lock detector output thus provides the lock control voltage that disables the VCLO sweep circuits and enables the range rate measurement loop. The loop detector output controls the VCLO and also demodulates any FSK subcarrier, range data tones, or speech subcarrier signals contained in the received signal. The output of this detector is distributed accordingly: (a) to the FSK subcarrier demodulating and decoding circuits, (b) to the range data filter-amplifiers, and (c) to the speech subcarrier demodulator. The range data filter-amplifiers separate the four tones and furnish the D1 through D4 signals to the range data translator circuits; also, a portion of each is properly phase-adjusted and the results recombined into a composite signal for use in phase-modulating the VCLO signal used to produce the first LO signal, thereby closing the "modulation feedback" loop.

d. Range Rate Data Logic. The range rate data handling section includes the $\dot{R}$ data trnaslator, a servo loop, and the range rate data and range rate reference counters. The f_o oscillator frequency (10.866 MHz) is binarily divided to $f_o/2048$ (5.306 kHz). The f_o output and four outputs of the binary divider, together with the output of the VCLO are applied to the $\dot{R}$ translator. By means of phase-lock loop techniques, the $\dot{R}$ translator then synthesizes the $\dot{R}$ servo frequency of $f_o/2048 \pm d/128$. The $\dot{R}$ servo loop thus is used as a tracking filter and doppler multiplier.

When the gates for the two 20-bit counters are enabled by $\dot{R}$ START, the $\dot{R}$ data counter counts the $f_o \pm 16d$ output of the servo loop VCO, and the $\dot{R}$ reference counter simultaneously counts f_o. If servo loop lock is maintained, any difference in the contents of the two counters at the end of the time period defined by $\dot{R}$ STOP is representative of the relative velocity of the interrogator with respect to the transponder.

The $\dot{R}$ servo loop includes the $\dot{R}$ LOCK detector whose output initiates a 30-ms delay, ensuring that the loop is settled before $\dot{R}$ START is issued. The $\dot{R}$ LOCK signal is also used in the control logic to provide the $\dot{R}$ LOSS memory bit in case lock is lost during a counting sequence (which invalidates the $\dot{R}$ data for that sequence).

e. Range Data Logic. The ranging frequencies are continuously generated in the frequency synthesizer which consists of two major circuit sections: a reference generator that produces the fine (FN), intermediate (INT), coarse (C), and very coarse (VC) range tones, and a modulation generator that produces the range modulation frequencies actually transmitted; i.e., to avoid using ranging tones within the capture range of the narrow-band loop, the intermediate, coarse, and very coarse range tones are folded about the fine range tone. The range modulation frequencies that make up the composite range modulation signal are designated D1', D2', D3', and D4', respectively, and the received range tones after separation are correspondingly designated D1, D2, D3, and D4.

At the appropriate point in each measurement sequence (following transponder verification and R LOCK delay) the composite range modulation signal is fed through a ramp generator and applied to the transmitter phase modulator. Over its 10-millisecond period the ramp generator linearly increases the modulation index from zero to full amplitude, providing final individual modulation indexes of 20, 2, 2, and 2, respectively, on the transmitter carrier.

The returned range modulation frequencies are coherently mixed in the range data translator with reference frequencies from the modulation generator so that each range modulation frequency is converted to a common servo frequency while retaining its phase information. The four outputs of the R data translator are fed to the range data servos where the phase of each is digitized in an 11-stage feedback counter. When a data sample is to be taken, each count is stopped at the zero crossing of the servo reference, automatically subtracting the reference from each range partial. The four 11-bit range word partials then are stored in their respective counters, ready for readout.

f. Voice Communications. The interrogator voice communication section provides voice communications with one or more of the transponders in the system. The voice communication subsystem employs a conventional headset unit. The push-to-talk switch on the headset microphone turns the transmitter on and off, and received audio signals are fed to the earphones on the headset.

g. Calibration Arrangements. For periodic calibration checks, the computer initiates a normal measurement sequence except that the calibrate command and dummy ID are issued in place of a valid transponder ID word, activating the calibration circuits. The calibration circuits then produce a 65.2-MHz (6 f_0) test signal and

mix it with a portion of the 1630.0-MHz (150 f_0) interrogator transmitter signal. The 1564.8-MHz (144 f_0) difference signal produced by the heterodyning action of the mixer is passed by the circulator and the receiving filter to the receiver circuits. The interrogator circuits acquire and lock to this simulated transponder signal and process the range and range rate data in the usual manner. The range rate data will indicate zero velocity. The range data resulting from each calibration sequence provides an updated calibration number representing the phase biases within the interrogator.

h. Self-Test Features. The CIRIS computer provides repeated operational checks and analyses of the interrogator alone (calibrate mode) and of the entire RRS (interrogator, antennas, cables, etc. plus transponder) during every normal interrogation sequence. The indications obtained permit a fault to be isolated to the line replaceable unit (LRU) level, and in most instances further isolates the trouble to the defective modular group or functional circuit area. Instead of relying on digital data smoothing of a large number of data samples to obtain reasonable accuracy, the RRS places emphasis on the accuracy of each data sample. This emphasis is maintained throughout the design of the equipment; e.g., all data channels are designed with more signal margin than the carrier channel, therefore if it is shown that carrier lock was obtained there is excellent assurance that the data is good.

In keeping with this design philosophy, the 7-bit data quality word not only detects the presence of data but also ensures that data errors will be minimized. In MSB-to-LSB order, each data quality bit is listed below and its significance is described.

(1) $\dot{R}$ LOCK bit: When a logic "1", this bit ensures that (a) carrier loop lock was achieved, and (b) range-rate servo loop lock was achieved and maintained long enough for the loop to settle.

(2) Malfunction bit: When a logic "0", this bit ensures that power supply voltages being monitored were present.

(3) R LOCK bit: When a logic "1", this bit ensures that (a) all four range data servo loops were phase-locked, and (b) lock was maintained throughout the settling time required for accurate range measurements.

(4) $\dot{R}$ LOSS bit: When a logic "0", this bit ensures that range-rate servo lock was maintained throughout the 900-ms $\dot{R}$ data measuring period.

(5) Verification bit: When a logic "1", this bit verifies that the ID of the transponder called up by the computer for interrogation and the ID of the transponder that responded correspond.

(6) Parity error bit: When a logic "0", this bit affirms that the transponder ID word supplied by the computer exhibited correct (even) parity. No transponder interrogation sequence can proceed until parity is obtained.

(7) FSK LOCK memory bit: When a logic "1", this bit assures that the FSK demodulator phase-lock loop acquired and locked (even though this loop was subsequently unlocked so that the ranging sequence could proceed).

To check against misleading data quality indicators, the computer can impose a signal dropout of controlled (10 ms) duration by interrupting the calibration control signal. The data quality word then is tested for proper detection of this signal loss.

Additional self-test features contained in the computer software program include overlap error tests E28, E38, and E48. For example, E28, is an 8-bit word that shows the amount of error in the INT range partial as compared to the FN range partial. The absolute magnitude of this word is a measure of data perturbations due to multipath, calibration error, noise transients, channel degradation, etc. At system threshold, the noise contribution error will be about 10 bits, whereas the ambiguity resolving algorithm corrects up to 128 bits (and has the capability to detect perturbations up to this limit without introducing range error). Thus, the magnitude of this error operates as both a short-term and long-term RRS data quality assessor, and assures 100 percent detection of bad data samples. In a similar way, the E38 and E48 words measure the amount of error in the INT/CS and CS/VC channels and offer the results for the evaluation of the computer. Further, the Kalman filter routine of the computer program tests the reasonableness of every range measurement, providing still another check.

4. <u>FUNCTIONAL OPERATION OF TRANSPONDER</u>: Figure 14 contains a simplified block diagram of the transponder circuits, and shows that the principal circuit sections include (a) a receiver and transmitter similar to those in the interrogator, and (b) an encode-decode (FSK SCO) circuit section for decoding the received ID word, and for reencoding and transmitting the verification ID word when the codes match. The transmitter carrier, first LO, second LO, and phase detector reference frequencies are all derived from a single voltage-controlled oscillator (VCO) that acquires and phase-locks to the interrogator carrier. After acquisition and phase-lock, phase coherence is maintained throughout the transponder in fixed relation to the received carrier frequency (e.g., the transponder transmitter carrier is maintained at a 24:25 ratio with respect to the received carrier), thus any doppler shift on the interrogator carrier is preserved. The following sequence of events describes the functional operation of a transponder.

- Unless responding to an interrogation, the transponder operates in a receive-only state and with its VCO (which generates the first LO signal) being swept ±30 kHz at a 40-Hz rate.

- The transponder receiver acquires and phase-locks to all interrogator transmissions received. The lock detector disables the VCO sweep generator, and the data detector produces an error voltage that causes the VCO to track the received carrier.

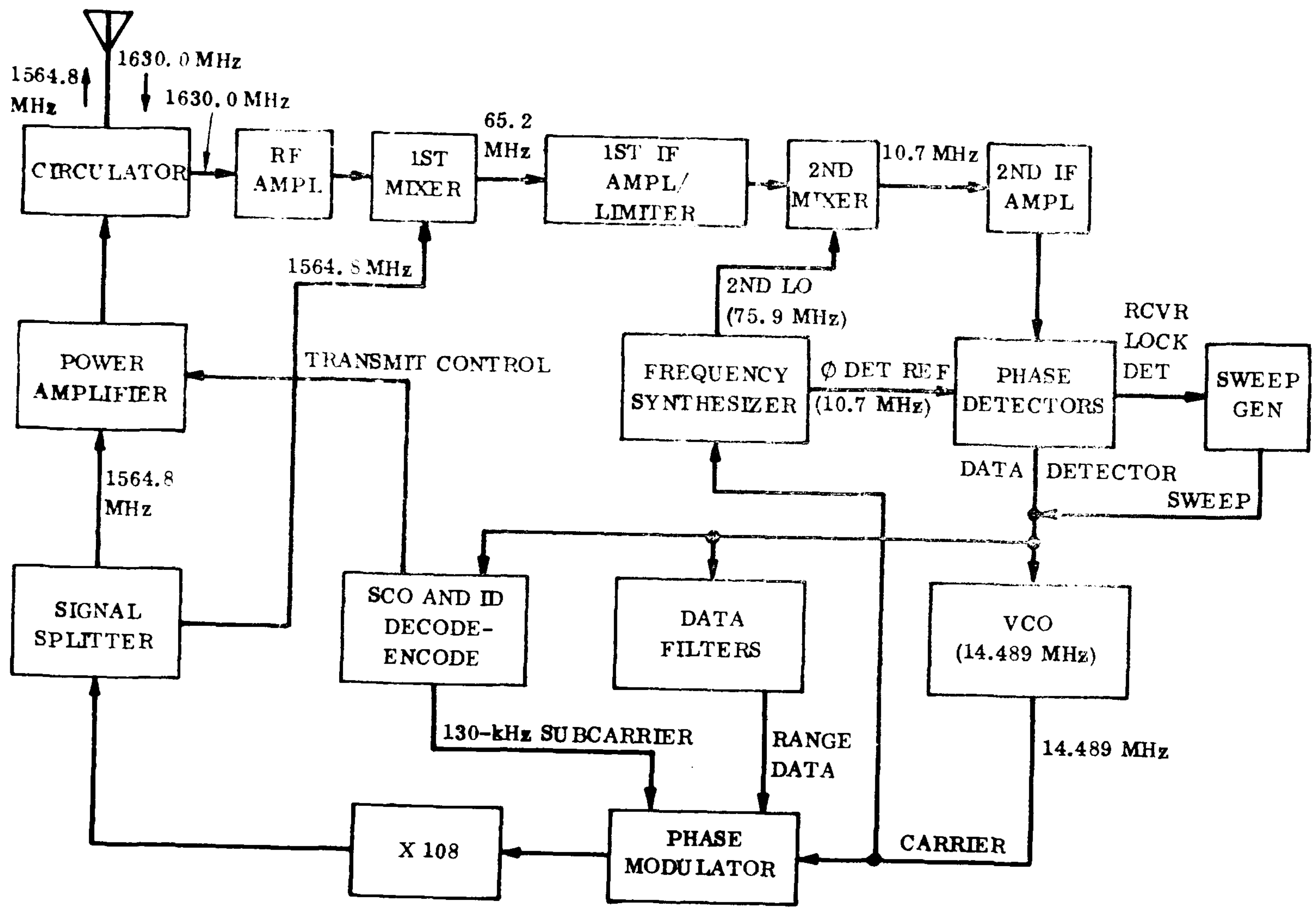


Figure 14. Simplified Block Diagram of Transponder Circuits

- The FSK demodulator phase locks to the FSK subcarrier and supplies FSK data to the slicer for amplification and shaping. The transponder programmer synchronizes to the received data pulses and furnishes shift and timing signals for data handling. FSK data bits then are shifted through a register where they are continuously compared with the transponder's ID code board. If no code match is obtained the transponder simply continues to track the carrier and FSK code until signal dropout occurs.

- In the event of code match, the transponder (a) turns on its transmitter, (b) serially recirculates and encodes its ID word plus sync, and (c) applies the resulting pulse train via shaping filter to a 130-kHz SCO. The FSK subcarrier is gated out to the transmitter modulator and thus is transmitted to the interrogator for a period of 35 ms. At the end of this period all FSK subcarrier circuits are deactivated.

- When the interrogator receives, demodulates, decodes and checks the transponder ID word sent via the 130-kHz subcarrier, it disables its own FSK SCO circuits. When range rate lock delay occurs, it modulates its carrier with the range modulation baseband.

- As range modulation appears on the transponder's received carrier it is demodulated, the range tones are separated and passed through a phase-adjusting network, then the range tones are recombined into a modulation baseband and applied to the phase modulator of the transponder transmitter. Since the transmitted signal is used also as the first LO, the same type of modulation feedback effect used in the interrogator is achieved in the transponder loop.

- The transponder retransmits the range modulation as long as it continues to receive the range-modulated carrier from the interrogator. When the interrogator stops transmitting, the transponder receiver breaks carrier loop lock, automatically turning off the transponder transmitter. The transponder then reverts to its receive-only state, ready for the next interrogation or voice communications.

a. Receiver-Transmitter Circuits. Although the transponder transmitter and receiver employ substantially the same circuit arrangements as the corresponding circuits in the interrogator, the transponder circuits differ in the following respects:

(1) Carrier and LO frequencies are derived from a 14.489-MHz voltage-controlled crystal oscillator (VCO) which, after carrier acquisition and phase lock, maintains the carrier frequency at a 25:24 ratio with respect to the received carrier frequency.

(2) The transponder receives on the 1630-MHz interrogator transmit frequency, and transmits on the 1564.8-MHz interrogator receive frequency.

(3) Since the transponder uses a low-level portion of the transmitter signal as the first LO, the first intermediate frequency is 1630.0 - 1564.8 MHz = 65.2 MHz. The frequency selected for the second if. is 10.7 MHz.

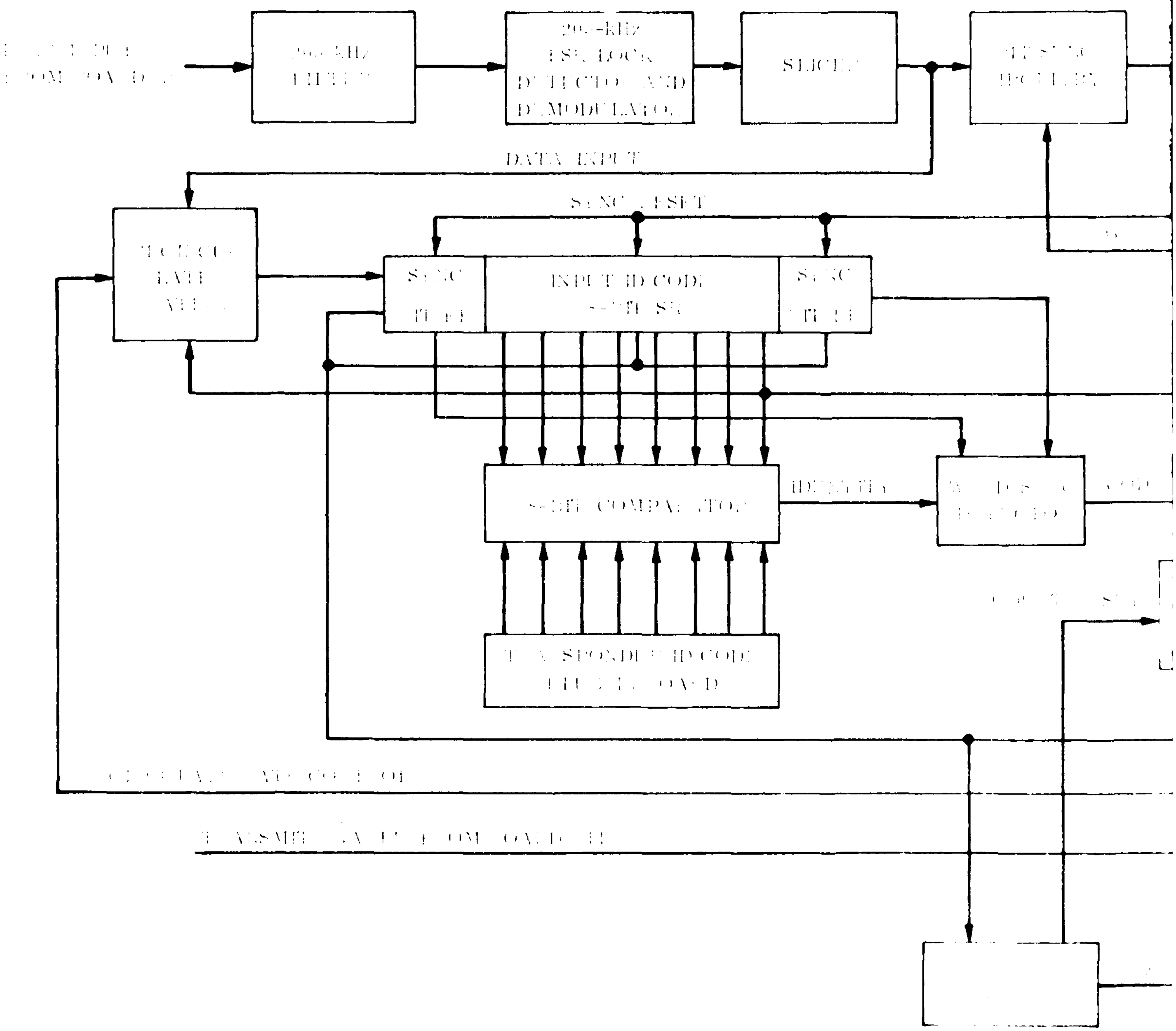
(4) For data handling, the transponder filters out and demodulates the interrogator's 205-kHz FSK subcarrier, and transmits the verification ID word by frequency-shift-keying a 130-kHz SCO.

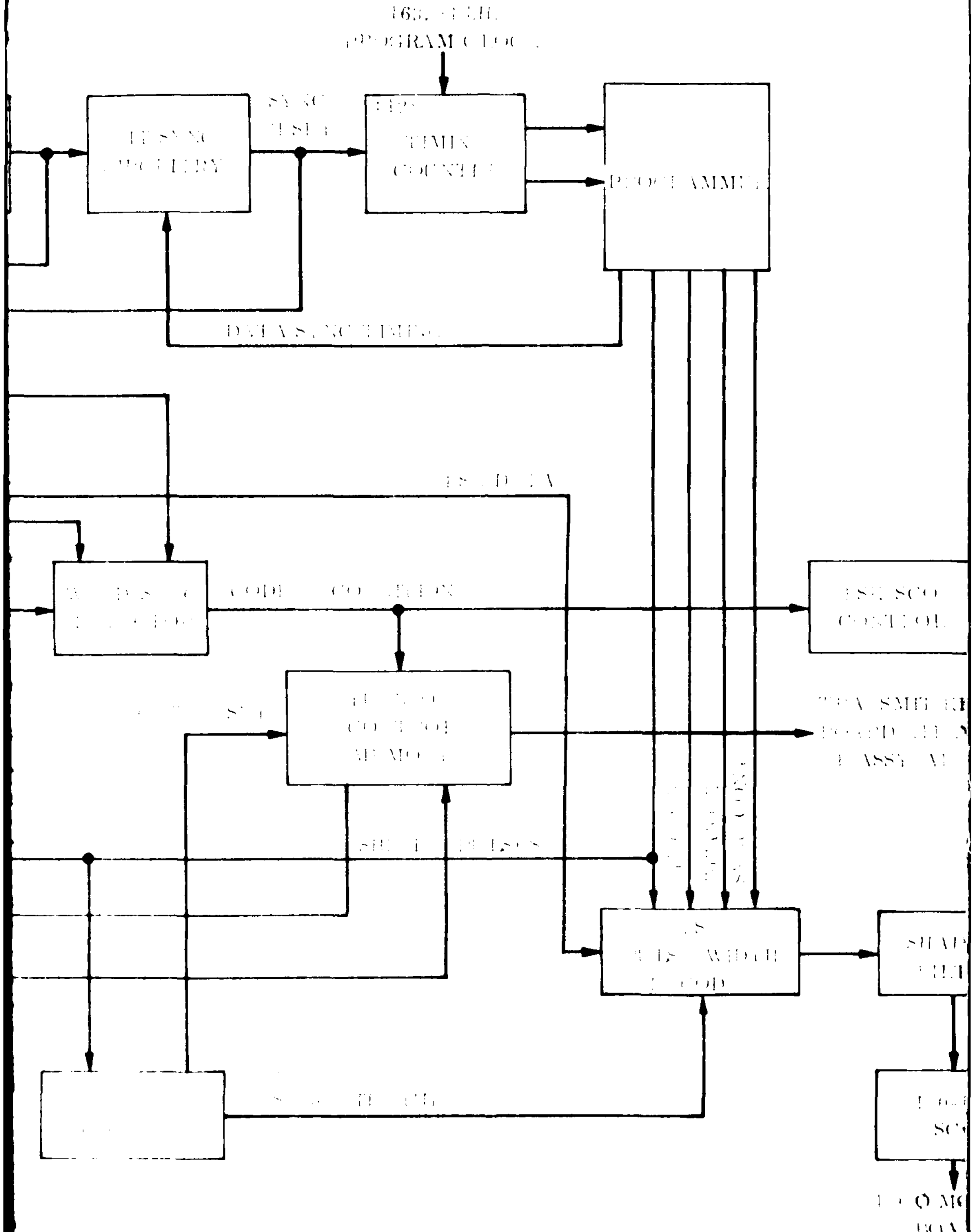
b. ID Word Decoder/Encoder and Control Circuits. At the start of every interrogation each transponder receives the 1630-MHz carrier modulated by a 205-kHz subcarrier which in turn has been FSK-modulated with the ID data word. This incoming signal is processed through the receiver and the decode-encode logic where the incoming data bits are compared with the unique hard-wired ID code matrix. If the two codes match, the transponder is turned on and the same ID word will be retransmitted to the interrogator as a verification signal.

Figure 15 contains a functional block diagram that shows how the 205-kHz ID subcarrier is processed and the 130-kHz FSK subcarrier is generated and supplied to the phase modulator/X18 multiplier. The receiver furnishes the 205-kHz subcarrier to board A10 where it is buffered by an emitter-follower stage. The 205-kHz FSK demodulator retrieves the ID data, and a slicer stage converts this signal to the original FSK pulse train (square-wave pulses). The ID data bits then go to board A11 where the demodulator and bit timing programmer operate in the same manner as the corresponding circuits in the interrogator.

The programmer provides properly timed shift pulses for shifting the decoded data through the shift register. To detect the ID word, the contents of the register are compared bit-for-bit with the hard-wired ID set into the plug-in ID code board. If code match and word sync are obtained, the code recognition signal is generated. This signal sets the flip-flop logic that provides the $\overline{\text{TRANSMIT KEY}}$ signal, turning on the transmitter. The transmitter then remains on until carrier loop lock in the receiver is lost.

The code recognition signal generates the FSK SCO power control signal which gates the output of the 130-kHz SCO through to the phase modulator/X18 multiplier. After ID code match is detected and the transmitter is turned on, the FSK loop programmer furnishes the control and timing pulses needed to gate the 8-bit ID word to the FSK SCO as the code is shifted out of its shift register. The 8-bit ID data gate output is pulse-width-encoded, word sync is added, and the entire pulse train is passed through a shaping filter to the 130-kHz SCO. Momentary loss of lock in the demodulator loop of 10 ms or less will not interrupt transmission of the ID verification word, but losses greater than 10 ms or bit timing lock loss due to disabling of the interrogator FSK SCO circuits will disable the gate controlled by the data phase lock signal.


DATA INPUT



F... are ... Functional ... of Transponder ID Decod... Concepts

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c. Voice Communication Circuits. The voice communication circuits in the transponder are similar to those of the interrogator. One difference is that various functions are controlled by gating circuit power through voltage regulators. Figure 16 is a simplified block diagram of the voice circuits.

Voice communication is initiated by closing the press-to-talk switch. This action enables a voltage regulator/gate device which provides +5V to inverter Q1, thereby energizing logic matching inverter Q2. The output of this stage is the "transmit enable" signal which connects to the transmit control logic. This logic produces two control signals: the transmit key signal that enables the +28V supply, and the voice power control signal that gates voltage regulator device on board A10. The voice power regulator supplies +10V to the 130-kHz SCO and the microphone amplifier. Since the transmitter is already turned on, the voice-modulated 130-kHz subcarrier phase-modulates the carrier and is transmitted to the interrogator.

If an audio subcarrier is present in the data detector output fed to the FSK and voice board it is demodulated by the 130-kHz demodulator, yielding the voice audio and a phase-lock indication. The audio is fed via the volume control and driver amplifier to the headset earphones. The voice phase-lock signal is used in the headset power control logic to generate the voice power control signal. This signal turns on a voltage regulator to power the headset driver stage.

d. Transponder BITE Circuits. The transponder is equipped with a press-to-test BITE switch that checks the operation of the transmitter and receiver circuits. When this switch is held closed the transmit key signal is generated, causing the transmitter to be turned on. With both transmitter and receiver energized, transponder operation is checked as follows:

(1) If the transmitter is radiating, a dc signal (transmitter BITE) is produced by the signal sampler in rf assembly A1.

(2) A receiver BITE signal (dc level) is generated, resulting in a calibrated noise level which denotes that the receiver circuits are operating properly. When both transmitter and receiver BITE signals are present they are ANDed, grounding one side of the BITE press-to-test lamp circuit and causing the lamp to light. This provides an indication that the overall operation of the transponder is normal.

5. INTERROGATOR TEST SET. Indicator-Interrogator Set Test Set TS-3301/URQ-32 can be used in place of the computer for controlling and checking the performance of the airborne interrogator. The test set thus is useful in conducting field shop checkout and maintenance of the interrogator, and can be used also to check the operation of a transponder connected to the interrogator by either cable or radio link.

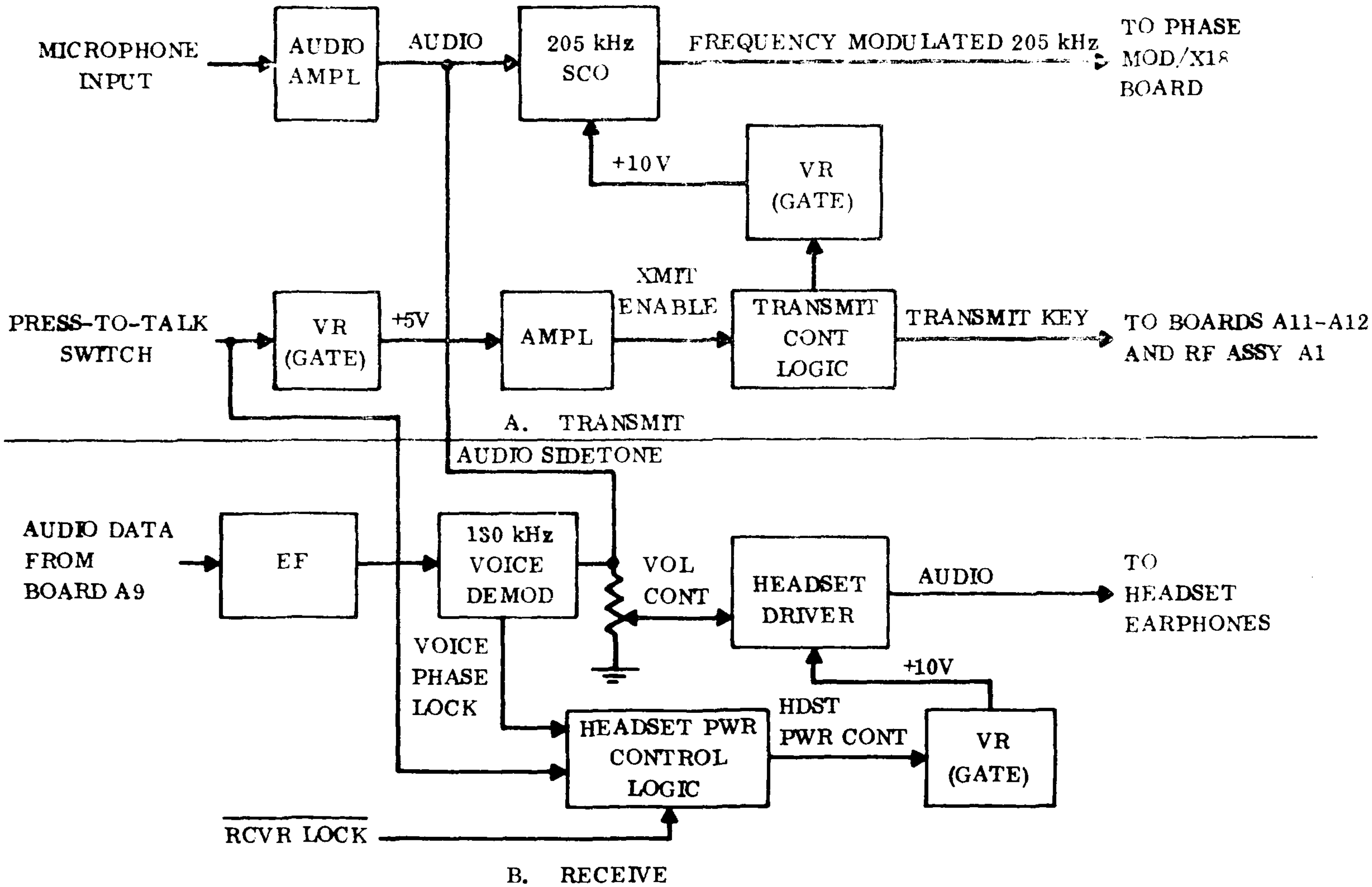


Figure 16. Simplified Block Diagram of Transponder Voice Circuits

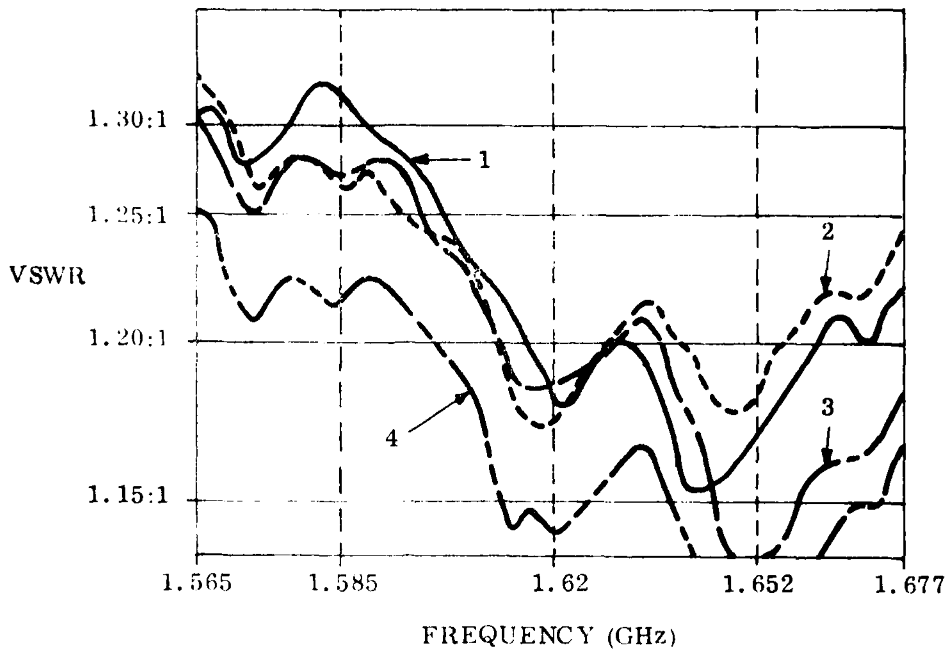
6. RRS ANTENNAS. One aircraft antenna and four ground station antennas were designed to meet system requirements, fabricated, qualification tested, and delivered with the equipment. The aircraft antenna, pictured in figure 2, is a quick-mount assembly consisting of a stub monopole, a thin-wall fiberglass radome, and 5-in. diameter mounting plate with standard Type N connector. The ground antenna assembly, pictured in figure 4, consists of a discone monopole, radome, and lightweight 4-ft. by 4-ft. ground plane that folds into a 2-ft. by 4-ft. package with carrying handle.

a. Design Approach. The RRS antenna requirements were analyzed to determine the optimum types to use for the airborne and ground antennas. Basically the requirement was for hemispherical coverage, specifying as end points a +3-dBi gain on the horizon and -23 dBi at zenith. An aircraft antenna used in previous Model CR-100 range/range rate systems was selected for the airborne application since its vertical stub operates in the frequency band of interest. The antenna type selected for the ground station was an omnidirectional antenna developed for use in the JIFDATS program, since particular care was taken in this design to maximize the gain on the horizon. The JIFDATS antenna was employed at Ku and C-bands, thus for the RRS the design was rescaled for operation at 1.63 GHz. Table 1 lists the performance characteristics expected of each type.

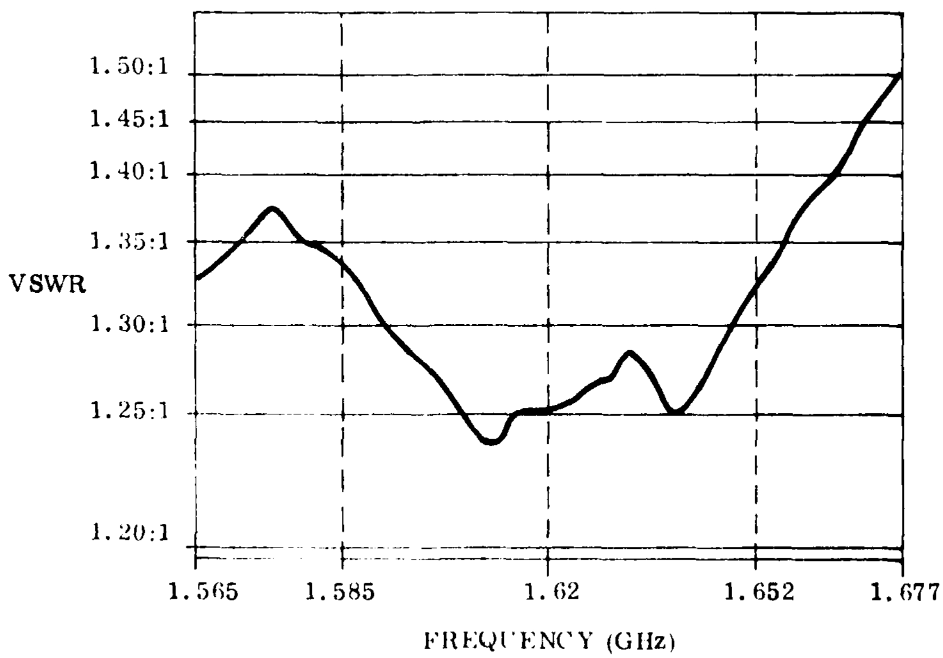
Table 1. Expected Performance Characteristics, RRS Antennas

Characteristic	Ground Antenna	Aircraft Antenna
Frequency Range	1.564 to 1.630 GHz	1.564 to 1.630 GHz
Impedance	50 ohms	50 ohms
VSWR (Maximum)	1.5:1	1.5:1
Gain at Maximum	+4.5 dBi (@ 27 ±3° above horizon	+4 dBi (@ 20 ±10°)
Gain on Horizon	-1.5 dBi typical	+0.5 dBi
Azimuth Pattern	Omnidirectional, ±2 dB	Omnidirectional, ±3 dB
Polar zation	Vertical	Vertical

b. Performance of Delivered Items. Tests conducted on the aircraft antenna and four ground antennas demonstrated that, when combined as a system, they will satisfy the RRS requirements for the CIRIS program. All data was taken with the antennas configured as they will be used in the RRS applications: transponder antennas in their radomes and installed on their 4-ft. square ground planes, the aircraft antenna installed on a ground plane 7 ft. in diameter to simulate an installation centered on the belly of an aircraft. The gain and VSWR characteristics of each antenna are tabulated below: (See VSWR plots vs. frequency, figure 17.)



A. TRANSPONDER ANTENNAS



B. INTERROGATOR ANTENNA

Figure 17. Plots of VSWR vs. Frequency, RRS Antennas

Antenna	Gain (dBi)	VSWR (Max)
Aircraft, Serial No. 1	+4.9	1.38:1
Transponder, Serial No. 1	+4.1	1.32:1
Transponder, Serial No. 2	+4.2	1.27:1
Transponder, Serial No. 3	+4.0	1.28:1
Transponder, Serial No. 4	+4.3	1.22:1

Figure 18 presents some typical azimuth patterns for the ground antenna, and figure 19 contains a typical elevation cut. Figure 20 shows the elevation pattern for the aircraft antenna, and figure 21 shows the combined or system gain, together with a plot of system requirements. The system gain pattern shows that the combined performance of the aircraft and ground antennas exceeds the system requirements by a large margin except for two narrow dips near zenith.

The area of reduced gain about zenith results from the necessarily small ground plane for the ground antenna. A larger ground plane would of course tend to fill this area and reduce the cone over which system requirements are marginal for worst-case conditions. (In the practical case, since the aircraft antenna's ground plane is not ideal its lobes tend to shift, therefore the "hole" also tends to shift and fill in, reducing the probability that the nulls of the two antennas will line up.)

In examining the two small areas where low signal levels may be encountered to assess their impact on system performance, it was noted that (a) for the cone at zenith the area is about ±10 degrees, thus at the maximum altitude of 50,000 feet the cone is 3.34 miles wide; and (b) for the horizon case the area 3 degrees above the horizon affects performance only at maximum ranges and reduced altitudes. For example, to range out to a distance of 200 miles on the earth's surface the aircraft must be at 27,000 feet altitude to be above the radio horizon for line-of-sight communication, corresponding to a grazing angle of 1.5 degrees. The combined patterns show the gain to be 2.3 dB below system needs at this point if ground station ground planes are horizontally leveled. As the aircraft closes the distance to the ground station the required gain decreases, so that at distances of 130 miles or less the system provides adequate reception without any range margin. It is important to note that the 3-degree point in question represents a range of 200 miles and an aircraft altitude of 50,000 feet, thus only at lower altitudes will performance be affected at maximum range.

It can therefore be concluded that, as shown in figure 21, the combined gain characteristics of the aircraft and ground station antennas greatly exceeed system requirements except for the cone around 10 degrees of zenith and at angles less than 3 degrees above the horizon at ranges beyond 130 miles. These factors are not considered

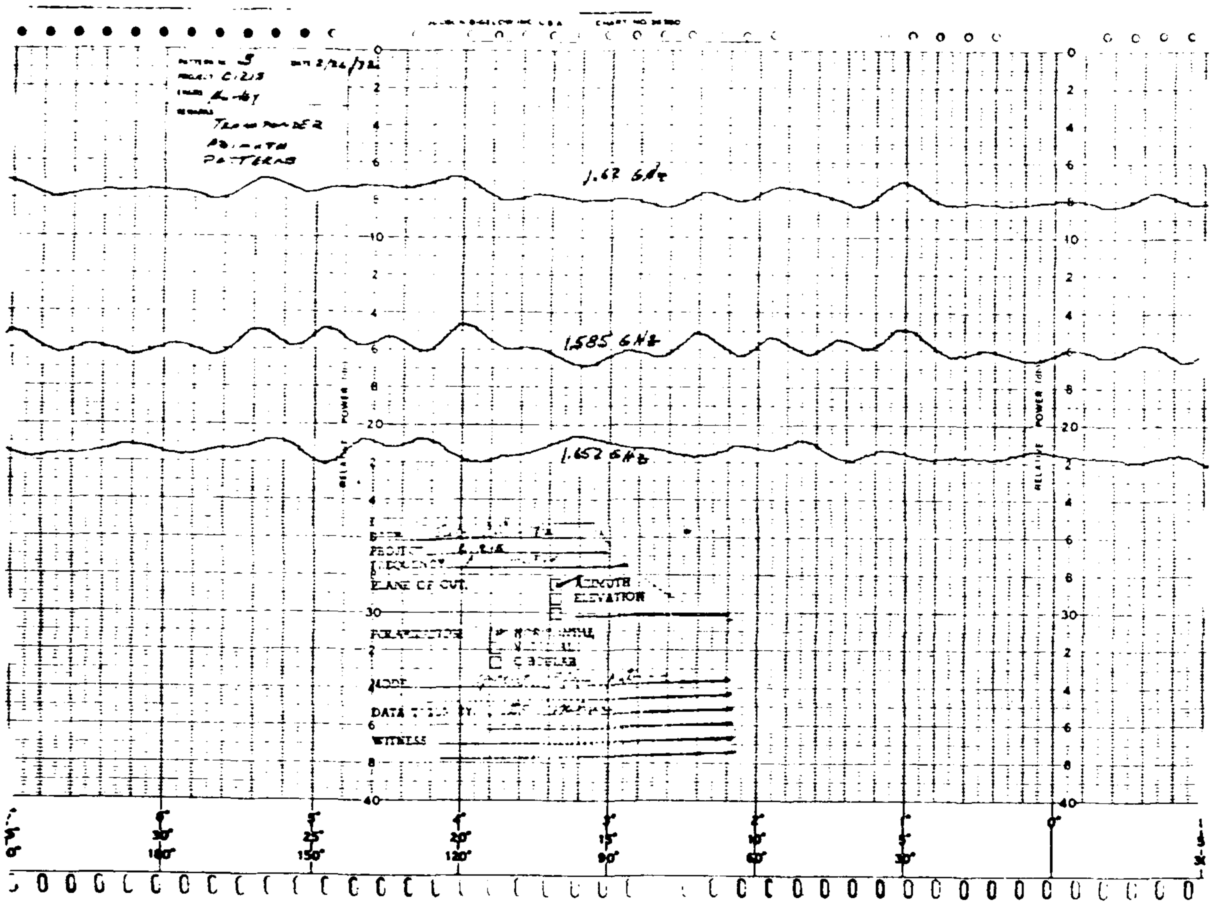


Figure 18. Typical Azimuth Patterns, RRS Ground Antenna

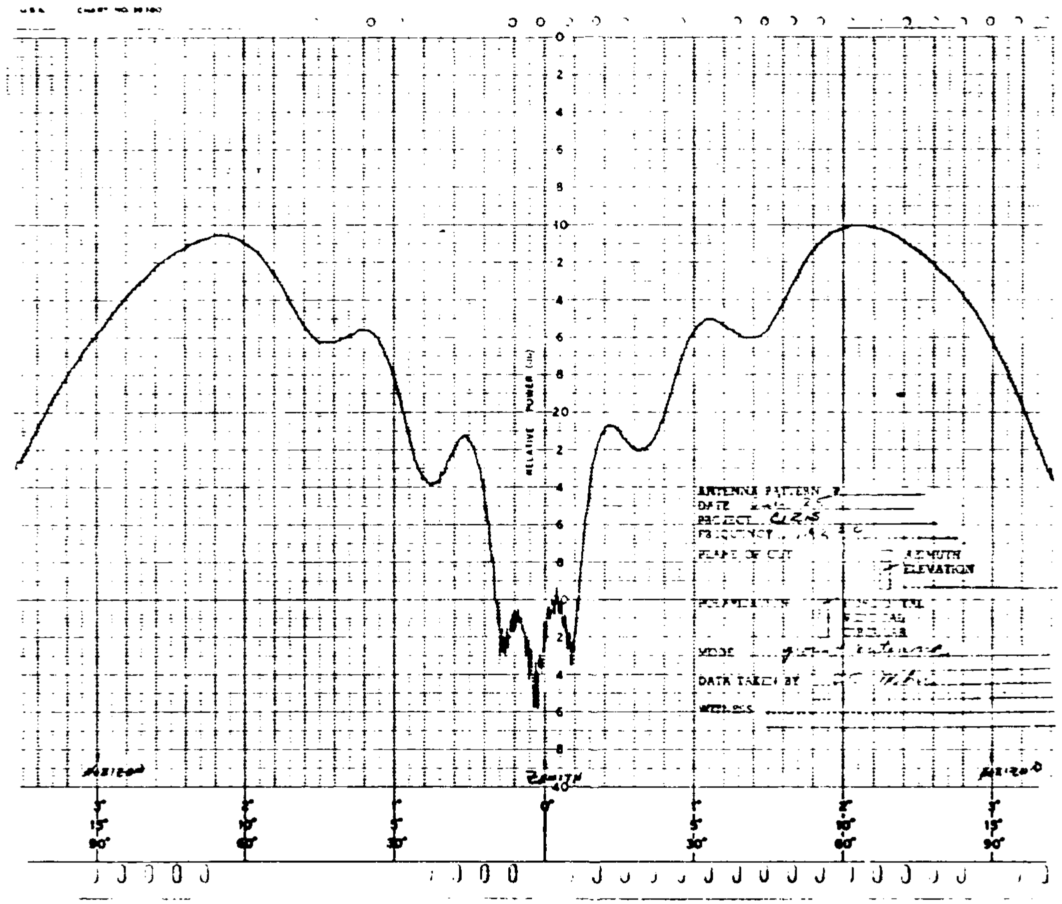


Figure 19. Typical Elevation Pattern, RRS Ground Antenna

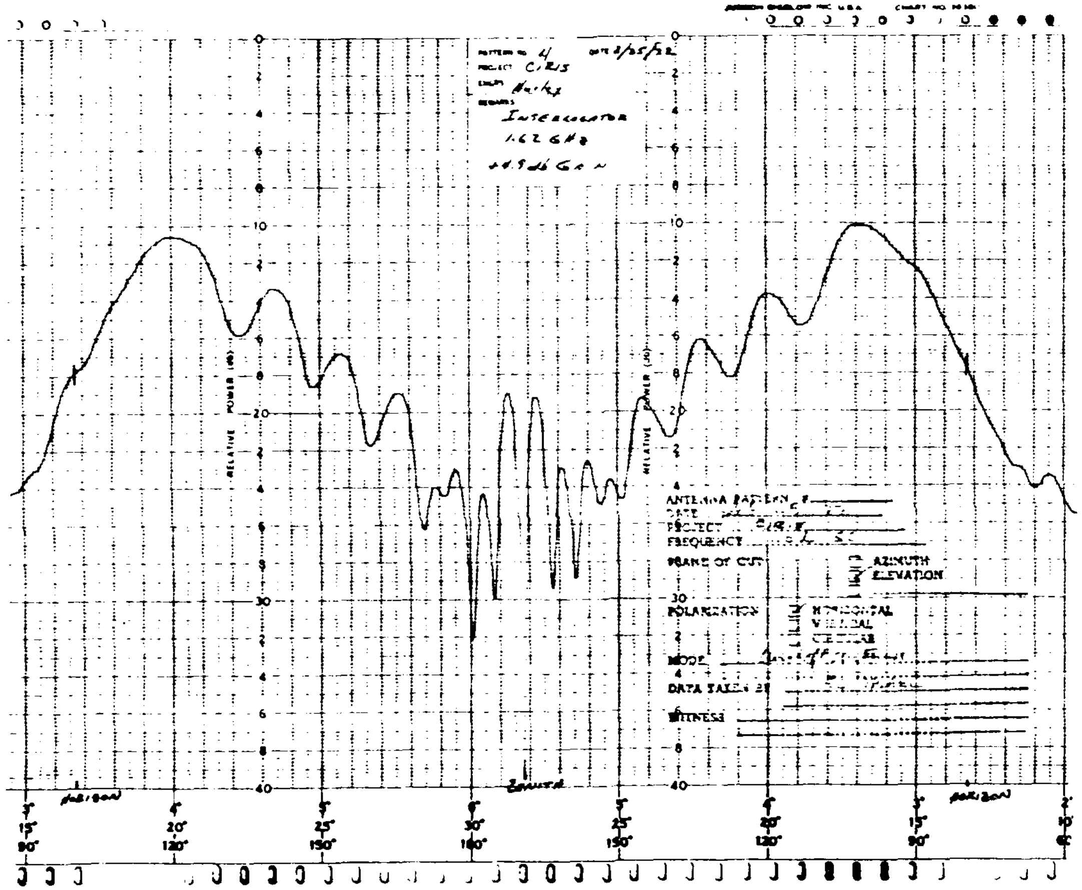


Figure 20. Elevation Pattern for RRS Aircraft Antenna at 1.62 GHz

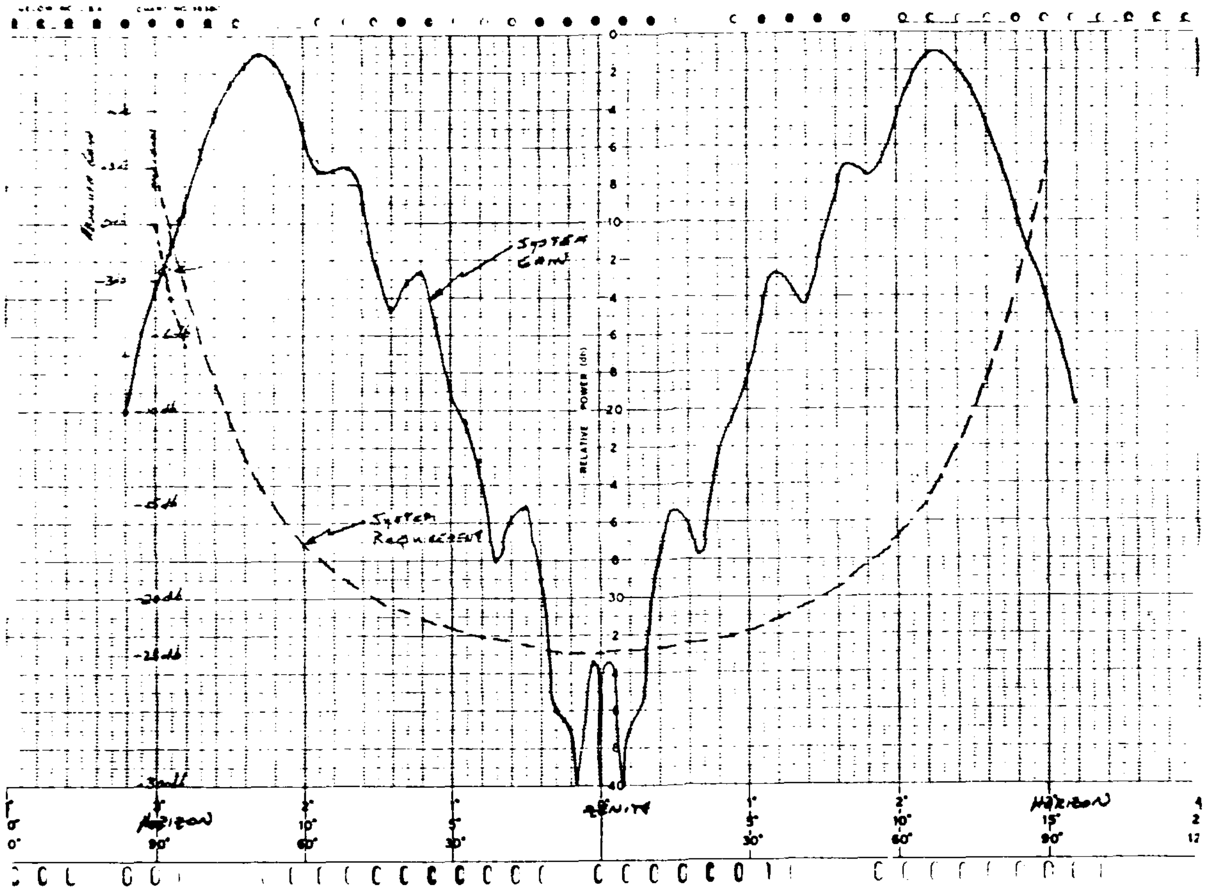


Figure 21. RRS System Antenna Gain vs. System Requirements

operationally significant since the area from the horizon to 3 degrees above the horizon is actually below the radio horizon, and the cone at zenith is relatively small. When a consistent flight pattern for CIRIS has been established, any areas of reduced gain at the horizon or zenith can be minimized or eliminated entirely by appropriately tilting the ground planes of the ground antennas to prevent the "holes" in the two patterns from lining up at zenith and, for the offset antenna, to favor low angle coverage. Figure 22 illustrates a typical arrangement for implementing this concept.

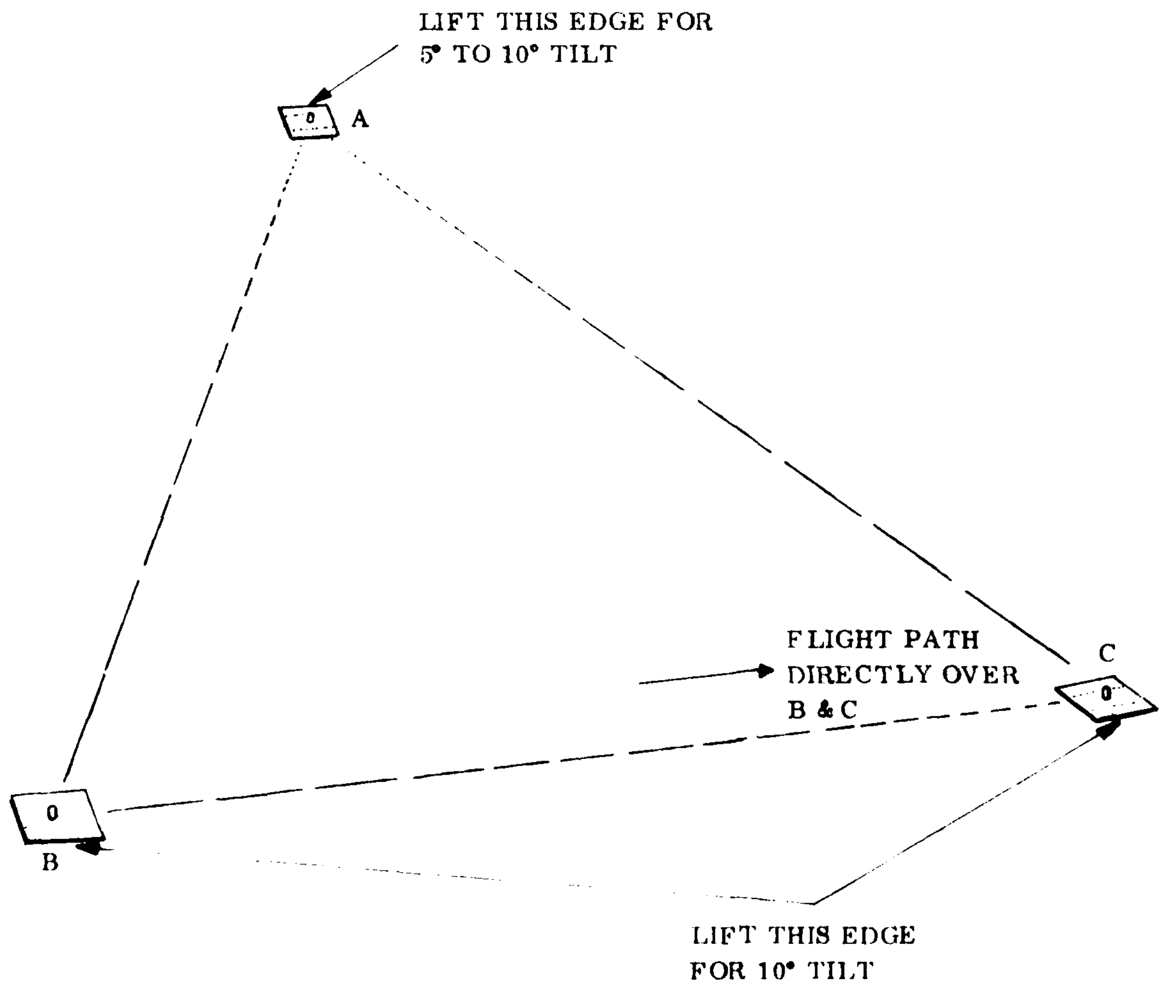


Figure 22. Typical Antenna Ground Plane Tilt for Operation with Overhead Flight Patterns

SECTION IV

COMPUTER/INTERROGATOR INTERFACE ARRANGEMENTS

1. INTRODUCTION. The Range/Range Rate Subsystem operates under the control of a general purpose digital computer, and requires the interface arrangements discussed generally in paragraph 2.c of section III. The following discussions provide specific computer/interrogator interfacing information in sufficient detail to enable the RRS to be interconnected with and operated by any suitable general purpose computer.

2. GENERAL REQUIREMENTS. The RRS interrogator is controlled by eight computer output command/control lines (or pairs) plus one optional interrogator power on-off control line, and supplies input data to the computer over 11 parallel lines (or pairs). The line receivers that respond to computer commands offer flexible input arrangements, accommodating either positive or negative logic and either floating-pair or one-side-grounded wiring. However, for this section the interface employed in connecting the interrogator to the Hewlett-Packard Model 2414B computer will be used as typical, wherein one side of each line pair is grounded to a common bus and negative computer logic is employed (ground or zero volts = logic "1", +5V dc = logic "0"). Although the interrogator itself employs positive logic, the interfacing logic elements provide any required logic level inversions.

3. COMPUTER COMMAND LINES. The computer controls the operation of the Range/Range Rate Subsystem by means of eight control pairs and one power control line, typically interconnected as shown in figure 23. These control lines are employed as follows:

a. The four address lines enter via J3 pins 1 through 8 and are applied to a binary-to-decimal decoder in the interrogator, yielding up to 16 decimal output lines. In the interrogator logic, these decoder outputs are designated address lines AD0 through AD15, where (a) AD1 through AD9 control the readout of interrogator output data, (b) AD10 and AD11 control the duration of the range rate measurement period, (c) AD15 issues a FIX command initiating a complete interrogation sequence, and (d) AD0, AD13 and AD14 are not presently used.

b. The strobe line (called "device command encode" in some computer arrangements) enters on J3 pins 21 and 22. The strobe operates as the enable/disable control for the address line decoder and calibrate command, and is used also to clock the transponder identification data word (8 bits plus parity) into the ID register and parity counter logic.

c. The transponder ID line enters on pins J3-17 and -18, and the ID gate enable is applied on pins J3-19 and -20. The enabling control permits the 8-bit ID word plus 1 parity bit to be serially strobed into the ID register and parity logic, then disables the ID gating logic during all other uses of the strobe.

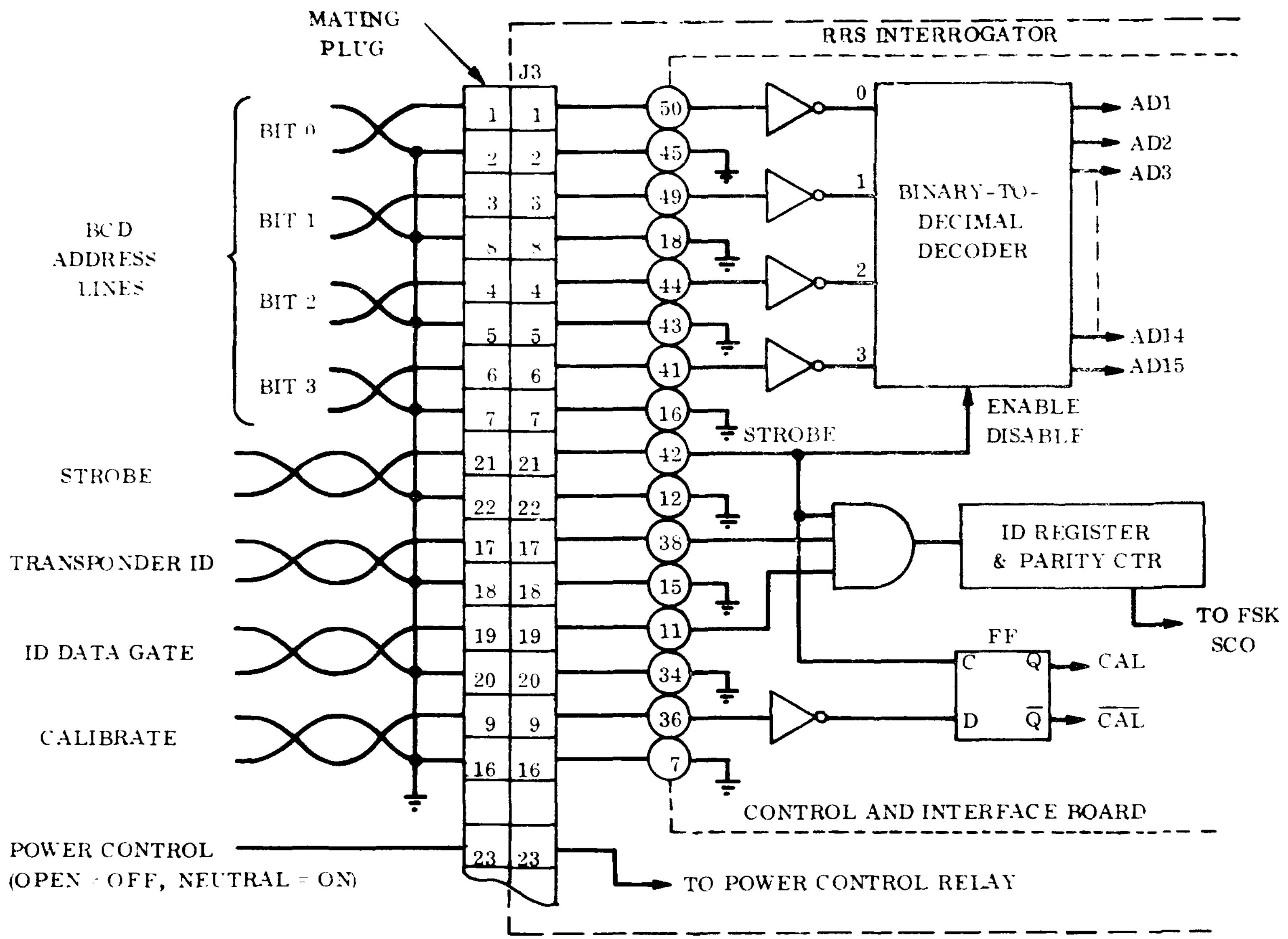


Figure 23. Interface Connections for Computer Command Lines

d. The calibrate command enters on J3-9 and -16, and is used by the computer to initiate an interrogator self-calibration sequence.

e. The power on-off relay can optionally be controlled by the computer by placing or removing a ground (or power line neutral return) on J3-23. Alternatively, the interrogator can be energized or deenergized using an ON-OFF toggle switch on the interrogator.

4. DATA TRANSFER LINES. The interrogator-to-computer data transfer lines connect via interrogator jack J3 as shown in figure 24. These consist of 11 parallel lines whose line drivers are driven via a wired-OR logic arrangement in the output data register logic. The data word placed onto the data transfer lines is controlled by the computer by means of address lines AD1 through AD9, where (a) AD1 through AD4 address the four range data words, (b) AD5 through AD8 address the range rate and reference words, and (c) AD9 places the data quality bits on the output lines. Table 2 lists the bit placements for each data word.

5. INTERFACE LOGIC ELEMENTS.

a. Line Receiver. The line receiver element shown in A, figure 25, is typical for all computer address line and command inputs to the interrogator interface. As this simplified schematic shows, the input circuit can be easily wired to accommodate floating two-wire inputs or a one-side-grounded arrangement of either polarity. The elements are designed for data rates up to 0.5 MHz, although faster rates can be used by removing the 1000-pf noise filtering capacitor at the input (at some risk of increasing the bit error probability).

b. Line Driver. The inverter logic arrangement shown in B, figure 25, is typical for each of the 11 data transfer lines. These can be easily wired to provide complementary outputs, or they can be used with one side grounded as shown.

6. SEQUENCING AND TIMING CONSIDERATIONS. The sequence of events in a typical RRS interrogation is illustrated in figure 11. To properly command the interrogator, monitor its responses; and transfer the resulting range, range rate, and data quality data to the computer; the software program must be arranged so that the computer performs the following tasks in the sequence given.

a. Using the transponder ID, ID data gate, and strobe lines, the computer serially shifts 9 bits (transponder ID word plus even parity) into the interrogator at the computer's data rate. (See figure 26 for typical timing diagram.)

b. At the time desired, start the interrogation by issuing the FIX command (AD15) using the four address lines plus strobe line. Note that the address line decoder logic responds to changes on the address lines only when the strobe line is in the

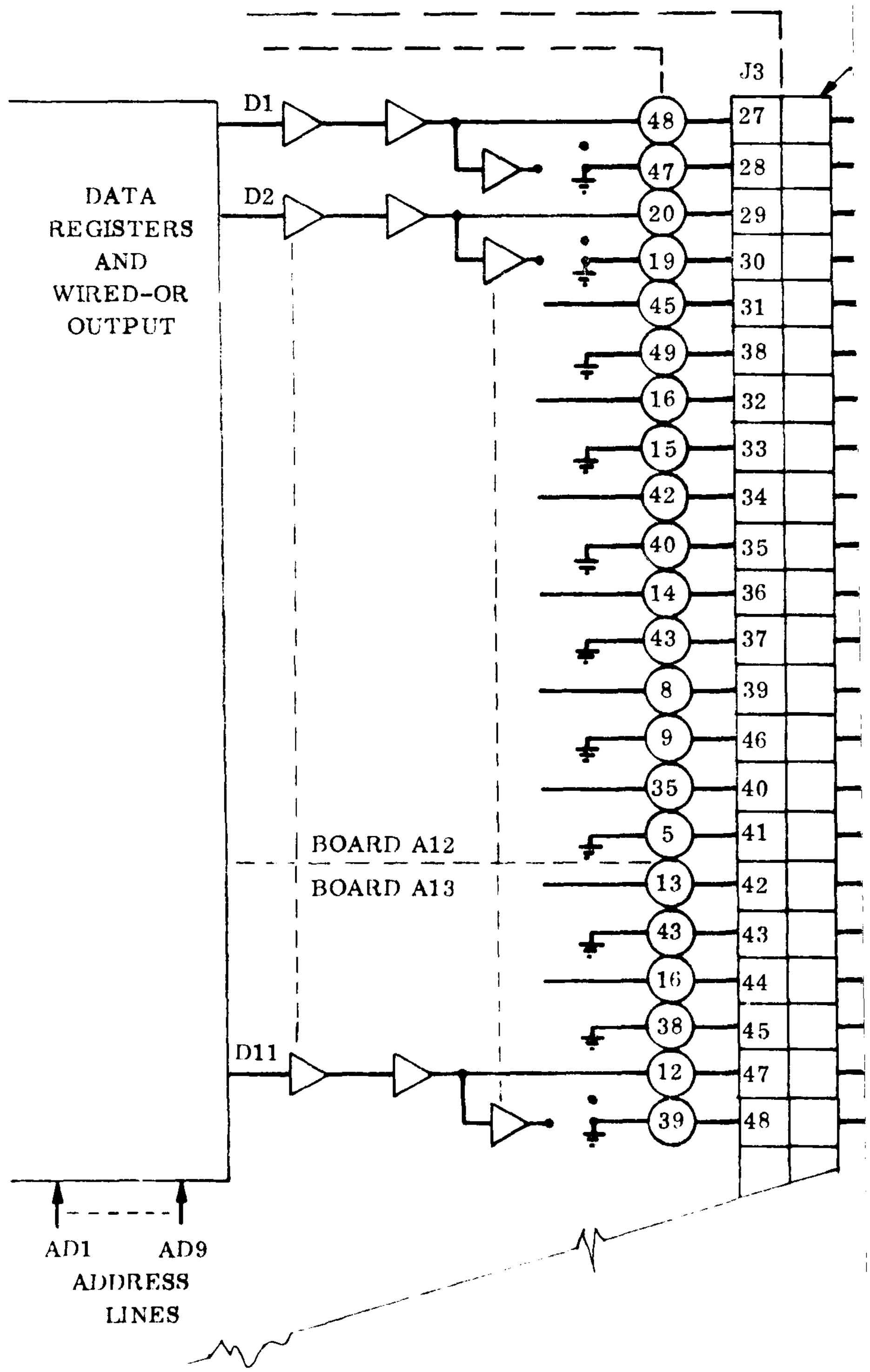
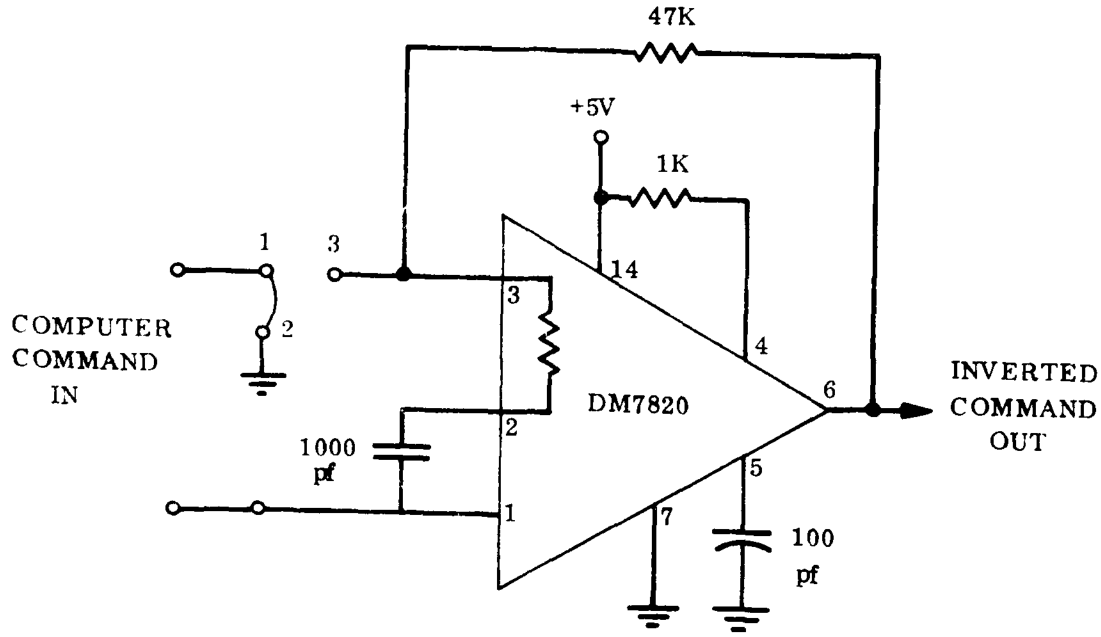


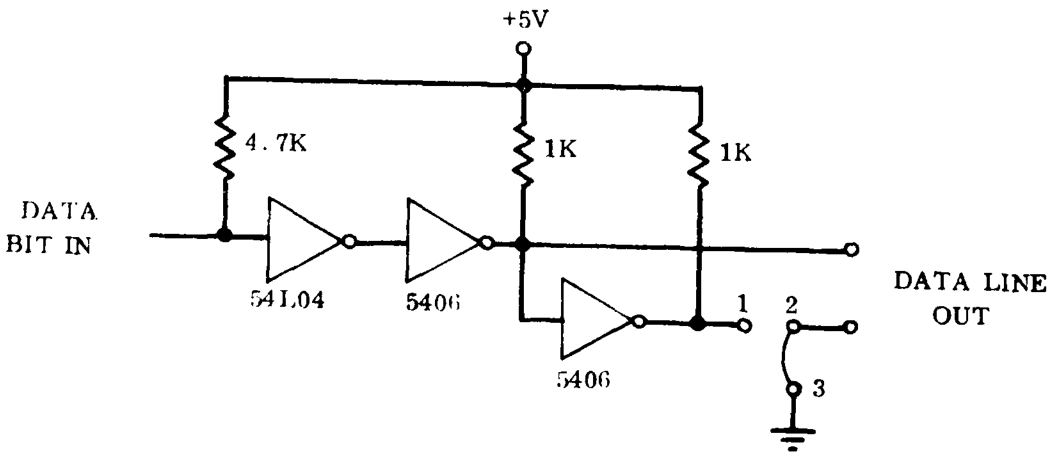
Figure 24. Interface Connections for Data Transfe

Table 2. Interrogator Data Words

Data Line	Data Readout Address No.								
	1	2	3	4	5	6	7	8	9
1	FN1	INT1	CS1	VC1	$\dot{R}1$	"0"	$\dot{R}$REF1	"0"	"0"
2	FN2	INT2	CS2	VC2	$\dot{R}2$	"0"	$\dot{R}$REF2	"0"	"0"
3	FN3	INT3	CS3	VC3	$\dot{R}3$	$\dot{R}12$	$\dot{R}$REF3	$\dot{R}$REF12	"0"
4	FN4	INT4	CS4	VC4	$\dot{R}4$	$\dot{R}13$	$\dot{R}$REF4	$\dot{R}$REF13	"0"
5	FN5	INT5	CS5	VC5	$\dot{R}5$	$\dot{R}14$	$\dot{R}$REF5	$\dot{R}$REF14	FSK LOCK
6	FN6	INT6	CS6	VC6	$\dot{R}6$	$\dot{R}15$	$\dot{R}$REF6	$\dot{R}$REF15	PARITY ERROR
7	FN7	INT7	CS7	VC7	$\dot{R}7$	$\dot{R}16$	$\dot{R}$REF7	$\dot{R}$REF16	VERIFICATION
8	FN8	INT8	CS8	VC8	$\dot{R}8$	$\dot{R}17$	$\dot{R}$REF8	$\dot{R}$REF17	$\dot{R}$ LOSS
9	FN9	INT9	CS9	VC9	$\dot{R}9$	$\dot{R}18$	$\dot{R}$REF9	$\dot{R}$REF18	R LOCK
10	FN10	INT10	CS10	VC10	$\dot{R}10$	$\dot{R}19$	$\dot{R}$REF10	$\dot{R}$REF19	MALFUNCTION
11	FN11	INT11	CS11	VC11	$\dot{R}11$	$\dot{R}20$	$\dot{R}$REF11	$\dot{R}$REF20	$\dot{R}$ LOCK



A. TYPICAL LINE RECEIVER



B. TYPICAL LINE DRIVER

Figure 25. Interrogator Interface Logic Elements

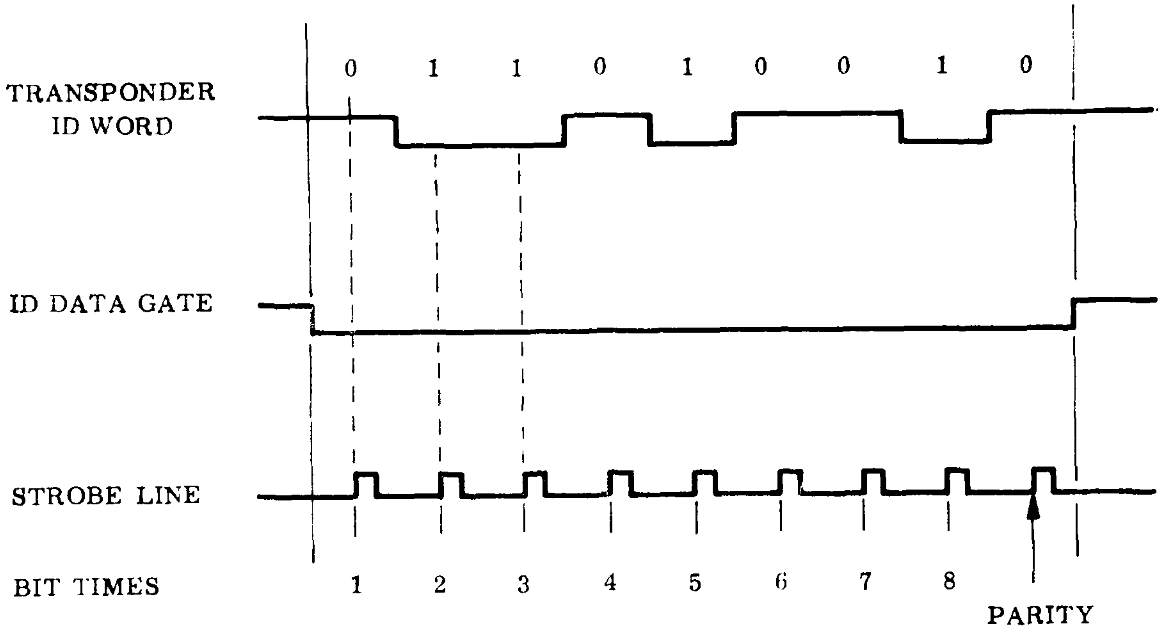


Figure 26. Typical Transponder ID Word Structure

logic "1" state; the address change then is accepted by the interrogator logic within 50 ns (gate delay time) after the strobe returns to the "0" state. Strobe width in either the "1" or the "0" state must be at least 1 microsecond.

c. Upon receipt of FIX, the interrogator first checks ID parity. If no parity error exists, it turns on the interrogator transmitter and enters "1" in the bit 6 location of the data quality (DQ) word. If parity is unsatisfactory, it sets a "0" in the DQ word and waits for computer to re-enter the transponder ID.

d. Thus, immediately after issuing FIX the computer should issue AD9 and monitor the DQ word (table 2). If parity checks, the computer's next concern is the presence of a "1" in bit location 7 of the DQ word, indicating ID verification; and a "1" in bit location 11, indicating range rate lockup ($\dot{R}$ LOCK).

e. $\dot{R}$ LOCK should be obtained within 110 ms after FIX and, when detected, the computer immediately issues $\dot{R}$ START (AD10) to start the range rate and range rate reference counters. If $\dot{R}$ LOCK is not detected within 110 ms, the computer must nevertheless issue $\dot{R}$ START and, after a brief interval, issue $\dot{R}$ STOP; this action aborts this particular interrogation and shuts down the interrogator transmitter.

f. For the normal interrogation, however, $\dot{R}$ LOCK will be obtained before the 110-ms deadline. The interrogator automatically conducts the ranging sequence, and if the interrogation is satisfactory in all respects it completes the DQ word so that:

(1) The first four (unused) bit locations will contain "0"s,

(2) FSK LOCK will be a "1" (locked obtained),

(3) PARITY ERROR will be a "0" (no error),

(4) VERIFICATION will be a "1" (verification obtained),

(5) $\dot{R}$ LOSS will be a "0" (no loss of carrier lock during R measurement),

(6) $\dot{R}$ LOCK will be a "1" (no loss of lock in ranging loops),

(7) MALFUNCTION will be a "0" (no malfunction), and

(8) $\dot{R}$ LOCK will be a "1" (carrier loop lock obtained).

g. Approximately 900 ms after issuing $\dot{R}$ START, the computer issues $\dot{R}$ STOP (AD11). The interrogator then completes the interrogation by (a) stopping the range rate and reference counters at the next zero crossing of the reference, then (b) shutting down the transmitter. Loss of received signal causes the transponder to shut down also.

7. DATA TRANSFER. The nine data words stored in the interrogator registers are available for transfer to the computer within 50 ns after the zero crossing of the range rate reference. The word structure for each readout address line is shown in table 2, wherein AD1 through AD4 transfer the four range data partials (fine, intermediate, coarse and very coarse), AD5 and AD6 transfer the range rate word, AD7 and AD8 transfer the range rate reference word, and AD9 transfers the DQ word. All range and range rate data words are in binary format, wherein bit location No. 1 is the least significant bit (LSB).

a. Sign Convention for Range Data. Each range partial (FN, INT, CS and VC) represents a complemented binary number in the existing unit. If the system line drivers were configured for differential outputs or the sense is reversed on the single-ended output, the range data would be available in positive uncomplemented form.

b. Sign Convention for Range Rate Data. The range rate counters count in a positive direction. When the motion of the interrogator is toward the transponder, for example, the range rate word will be larger than the reference, therefore R - RR will yield a positive binary number.

8. RANGE CALIBRATION. Since range is measured by measuring the round-trip phase delay of the modulation baseband from interrogator antenna to transponder and back, all phase delays due to signal paths within the equipment must be accounted for. For each transponder, for example, a calibration is conducted and the resulting calibration data is entered into the computer software to be applied as a correction to the final range word. For the interrogator, some of the circuitry is not within a feedback loop but is included in an internal calibration loop, therefore a method for frequent re-calibration during CIRIS missions is provided.

a. Calibration Frequency. Factors affecting how often the interrogator should be calibrated include (a) time available in operating scenario and (b) the predicted severity of the operating environment. Ideally, a calibration sequence would precede each interrogation. However, except in extreme temperature variation conditions an interrogator programmed for every 10 or 20 samples provides satisfactory accuracies. Further, if operating in a controlled, air-conditioned environment, after warmup to operating temperature a frequency as low as every 5 minutes could be adequate.

b. Calibration Sequence. For range calibration sequences the ID word entry is omitted, thus the sequence begins with (a) setting the calibration line to a "1", (b) issuing a FIX command (command address AD15), and (c) strobing the address line, which also sets the calibration flip-flop. (See figure 23) The interrogator then conducts a self-interrogation that differs from a transponder interrogation in the following respects:

(1) The received signal is a simulated transponder transmission originating within the interrogator.

(2) Since the interrogator transmitter signal is received and processed by the interrogator receiver, the four range data words represent the phase delay introduced by the interrogator. Following the issuance of $\dot{R}$ STOP these words can be transferred to the computer in the usual manner but stored for use in correcting operational range data.

(3) The range rate measurement should yield an $\dot{R} - \dot{R}R$ difference of zero, thereby serving as an operational check of the range rate measuring capability. In addition, a check on the validity of the R LOSS data quality bit can be conducted by (a) setting the calibrate line to "0" for a period of 10 ms or longer sometime after R START is issued, (b) setting all "0"s on the address lines for the period between R START and R STOP and (c) strobing the calibration data out and back in.

SECTION V

NEW CIRCUIT DEVELOPMENTS

1. INTRODUCTION. The following paragraphs describe some of the circuits that were developed in connection with the design and development of the RRS equipment units.

2. MICROSTRIPLINE PREAMPLIFIER. The new preamplifier is a low-noise unit employing a microstripline design on a Teflon substrate. Its function in the receiver is to amplify received microwave signals before any mixing or signal conversion takes place. Since this unit establishes the noise figure for the receiving system, it also determines the receiver's sensitivity.

To perform efficiently the preamplifier must exhibit superior gain and stability characteristics under all operating conditions likely to be encountered. Experience with the more conventional stub-tuned interstage coupling methods indicated that such approaches do not offer the required stability because the bandwidth yield of stub tuning is too narrow. Accordingly, as soon as the transistors to be used were selected (NEC 1336 input stage, HP 35821E second stage), a new form of broadband matching network was devised for the input and interstage coupling. These coupling arrangements took the form of exponential horns, as illustrated in figure 27, and proved to provide a 40-percent bandwidth. Also, because the horns are physically small they simplified the packaging problem, easily permitting the in-line layout that is virtually decreed for microwave amplifier designs to minimize feedback. Additional design features include the following:

a. A short-step Chebychev transformer provides the output coupling. The bandwidth is quite wide, and this device offers an ideal match between the collector of the output stage and the image rejection filter that follows.

b. The preamplifier is tuned by means of two small (0.3 to 3.5pF) variable capacitors. C3, placed at the input rf port, is tuned for minimum noise figure at the frequency of interest; and C13, at the rf out terminal, is tuned for an output VSWR of 1.5:1 or less when the preamplifier input is terminated in a 50-ohm load. To help stabilize the preamplifier when connected to its image rejection filter, 68-ohm resistor R7 was added to the rf output port.

c. Since the transistors are grounded-emitter types, collector-stabilized dc biasing is employed, with the first stage biased for low-noise operation and the second stage biased for best gain.

Early models of this preamplifier design exhibited the following characteristics:

	1.5648 GHz	1.6300 GHz
Gain	18 dB	17.5 dB
Noise figure (average)	4.5 dB	4.5 dB
Bandwidth (minimum)	200 MHz	200 MHz
Input VSWR	1.25:1	1.04:1
Output VSWR	1.15:1	1.12:1
Total power (@ 15V)	300 mW	300 mW

Stability was excellent under all conditions, the 1-dB gain compression point occurred at approximately -16 dBm, and the gain variation over the full temperature range was less than 2 dB. In the production units used in the RRS equipment, the gain was increased to 20 dB by replacing the second-stage transistor with a type HP 35861E. All other operational parameters cited above remained valid.

3. TRANSMITTER POWER AMPLIFIER. An output power amplifier satisfying a number of design criteria was required for both interrogator and transponder transmitters. The criteria included (a) an output of 6 watts or more at the carrier frequencies (1.5 - 1.6 GHz region), (b) a flat (1-dB) pass band of 15 MHz or more, (c) constant phase delay over the operating temperature range to avoid introducing error into the range measurements, and (d) the repeatability characteristics, reasonable cost, and minimal complexity essential for cost-effective quantity production.

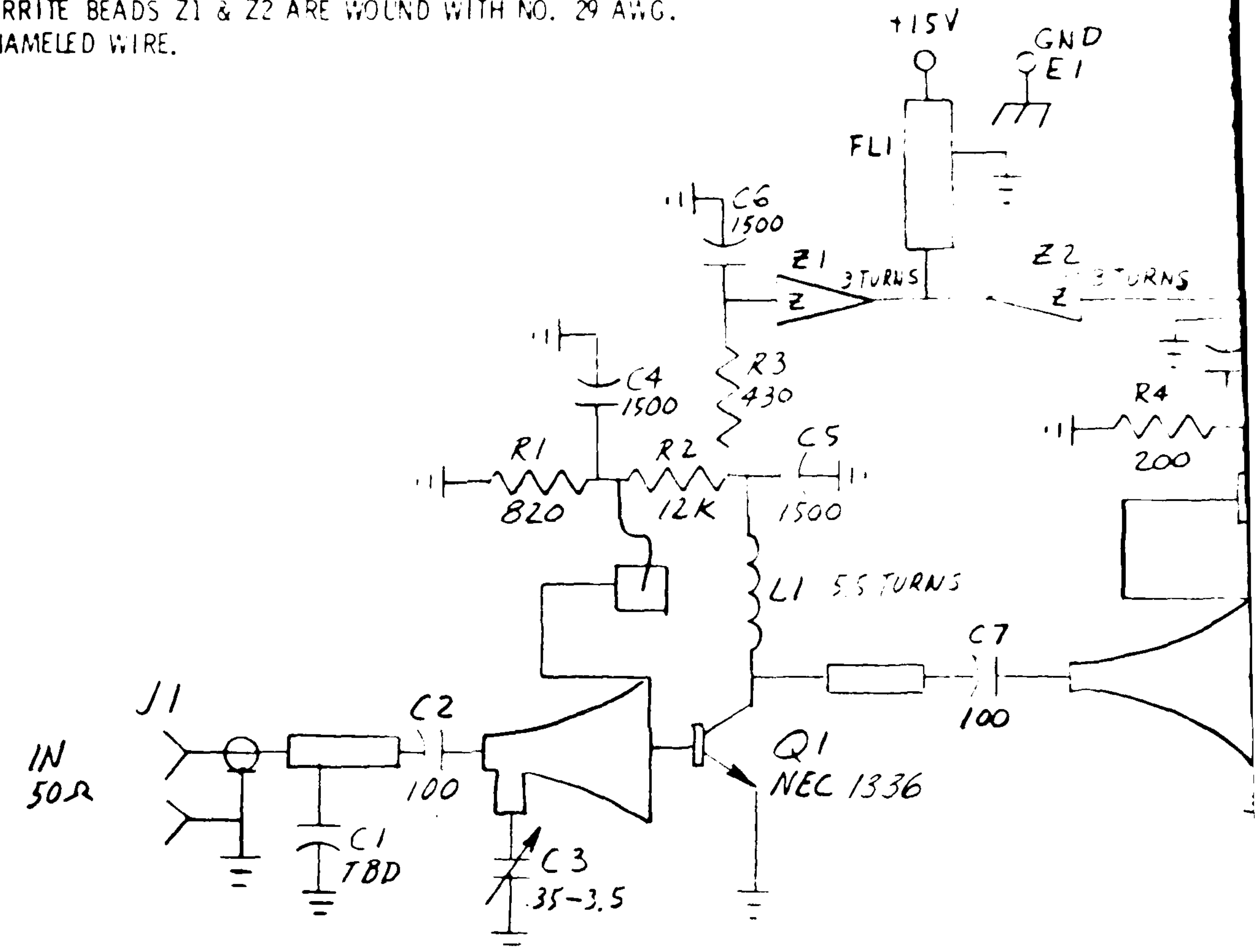
For the frequency range demanded for the amplifier, three design approaches were considered: lumped-constant (discrete inductors and capacitors), cavity-tuned, and microstripline. Lumped-constant methods offer the greatest flexibility (particularly in the breadboard phase), are simple and easily understood, and thereby constitute a popular technique; however the components get quite small at gigahertz frequencies. Tuned cavities are predictable and work well, but are cumbersome, inflexible, and awkward to fit into the simple packaging concept desired for the RRS. Microstripline techniques appeared to offer an optimum compromise with respect to the other two methods, easily meet the small size requirement but are somewhat difficult to design.

Both lumped-constant and microstrip designs were tried, and the microstrip unit was found to be superior. The final design, illustrated in figures 28 and 29, employs two Microwave Semiconductor Corporation transistors: an MSC 2001 used as a 1-watt driver, and an MSC 2005 used as the 7.5-watt output amplifier. Substantially identical circuit arrangements are employed in both stages.

Basically, designing a microstrip amplifier requires that both input and output impedances be properly matched to the source and output device, respectively. Using a

NOTES: UNLESS OTHERWISE SPECIFIED

1. RESISTANCE VALUES ARE IN OHMS.
2. RESISTORS ARE ¼ W.
3. CAPACITANCE VALUES ARE IN PICO FARADS.
4. FERRITE BEADS Z1 & Z2 ARE WOUND WITH NO. 29 AWG. ENAMELED WIRE.



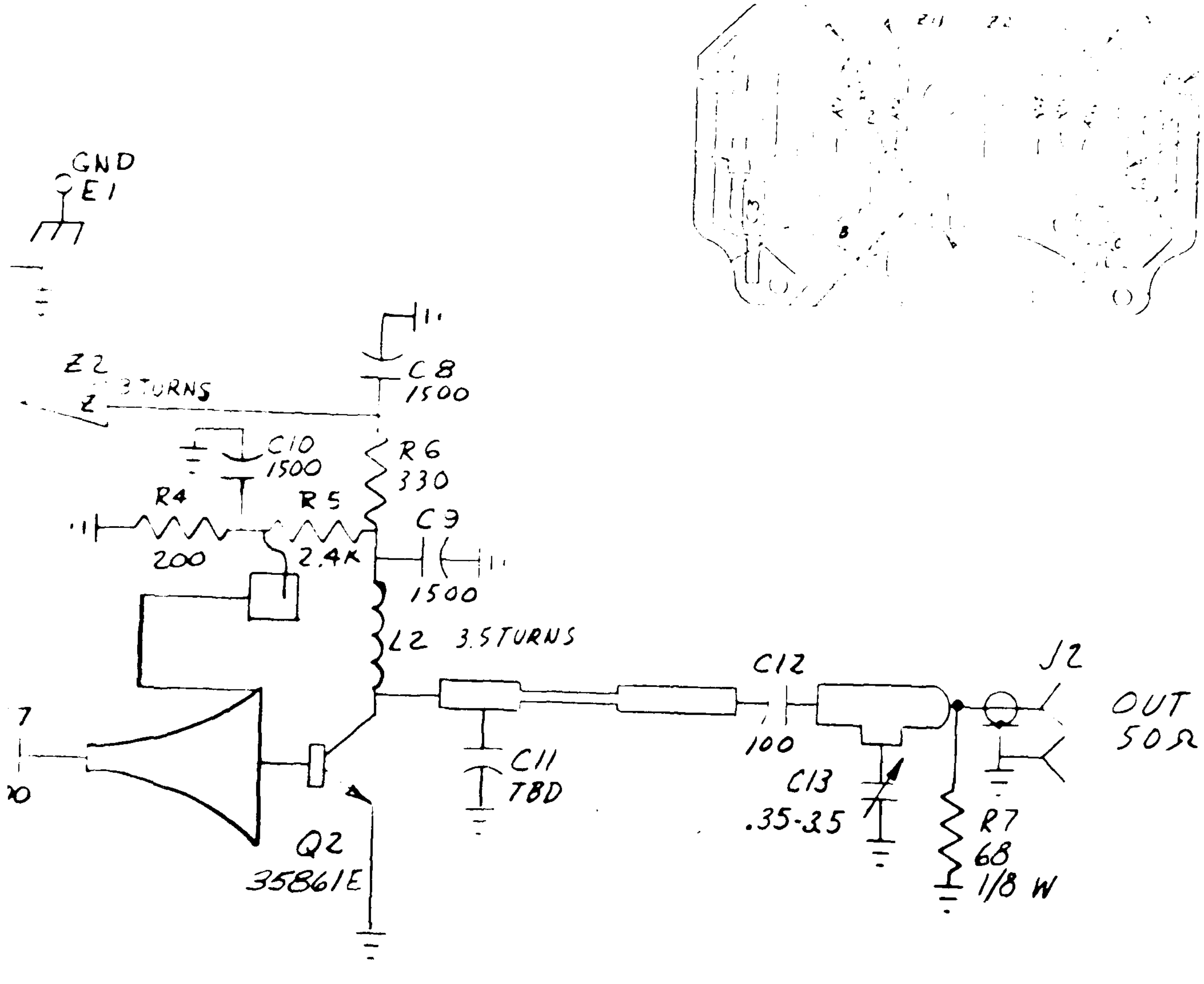


Figure 27. Microstripline Preamplifier

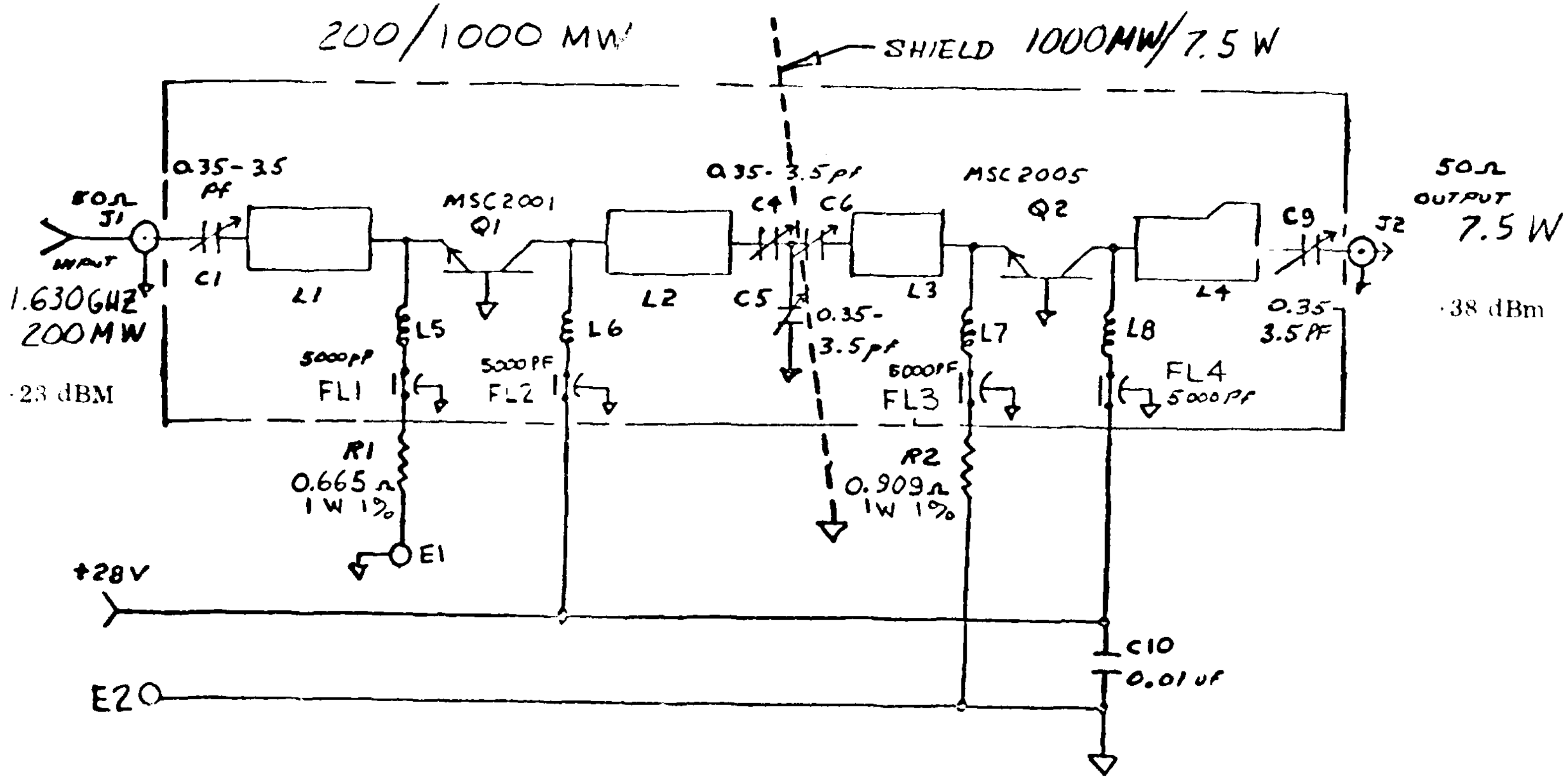


Figure 28. Schematic Diagram of RRS Power Amplifier

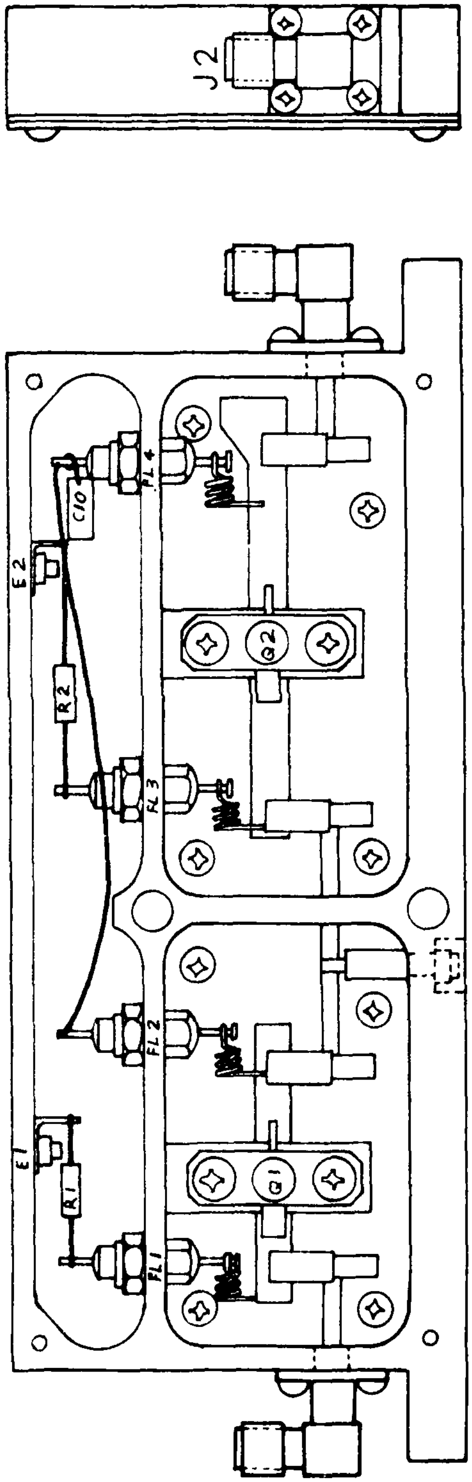


Figure 29. Assembly Drawing of RRS Power Amplifier

128111-E

Smith chart and the transistor manufacturer's design information, a 50-ohm input network was devised using a capacitor in series with a 15-ohm microstrip transmission line of suitable length, providing the required input match. The capacitor provided the necessary dc blocking, and a choke and decoupling filter provided power supply isolation. Similarly, 50-ohm output matching was obtained using a 30-ohm microstrip of proper length in series with another capacitor. The resulting amplifier circuitry met all design criteria, showed no tendency to oscillate, was easily aligned, and provided a 40-MHz bandwidth at the 1-dB points, greatly exceeding the 10- to 15-MHz needed for good phase stability.

Although the driver and PA stages share a common housing they are essentially independent amplifiers well shielded from each other. Proper operation requires fabrication techniques that will ensure that an effective, unvarying ground plane be maintained. Therefore screws are used to hold the circuit board in snug contact with the cast aluminum housing, and additional screws and Teflon spacers are used to hold each transistor firmly in place. These features provide the uniform ground plane needed for stable operation and also enable the module housing to act as an effective heat sink for the transistors. Output power proved to be 7-1/2 watts nominal (but 8 watts typical). When changing frequencies, emitter resistors R1 and R2 can easily be selected to keep the output power within the maximum ratings specified by the manufacturer for the frequency in use.

4. IMPROVED OSCILLATOR CIRCUITS. Two fixed-frequency crystal-controlled oscillators and two voltage-controlled crystal oscillators (VCXO's) are employed in the RRS as follows:

Cubic P/N	Frequency (MHz)	Use
128326	1.920423	Used to synthesize all ranging tones
128327	10.866667	Multiplied by 150 to generate interrogator carrier frequency
128325	11.673177	VCXO for interrogator receiver phase-lock loop; multiplied by 128 to produce first LO frequency.
128324	14.488889	VCXO for transponder receiver phase-lock loop and transmitter carrier; multiplied by 108 to generate first LO and carrier frequency

For the RRS to obtain accurate range and range rate data these four oscillators must be frequency-stable with time and over a wide temperature range. To conserve space and input power, temperature-compensated oscillator circuits were to be used in lieu of a temperature-controlled oscillator oven.

a. Statement of Problem. Earlier models of the Cubic CR-100 Range/Range Rate Subsystem (upon which the CIRIS RRS is based) experienced some oscillator stability problems. For example, voltage-controlled oscillators employed in the carrier tracking loops tended to drift from center frequency with time and temperature, causing acquisition problems. Since system accuracy in both range and range rate measurements depends upon stable fixed-frequency crystal oscillators and dependable VCO's, a product improvement program was conducted for the CIRIS RRS oscillators.

b. Performance Specifications. To initiate procurement of temperature-compensated oscillator modules meeting the demanding performance characteristics required, specifications were drawn up. Units meeting these specifications were procured from a leading specialist in this field (Arvin Frequency Devices). For example, the electrical characteristics specified for fixed-frequency crystal oscillator 128327 and VCXO 128324 included the following:

Characteristic	XOSC 128327	VCXO 128324
Frequency	10.866667 MHz	14.488889 MHz (6.20V dc on control line)
Frequency adjustment range (min.)	±6 ppm	±6 ppm
Stability over temp. range specified	±1 ppm, −40 to +80°C	±10 ppm, −20 to +80°C
Stability:		
Short term	1×10^{-8}/sec, average period	1×10^{-8}/sec, average period
Long term	1×10^{-6}/year	1×10^{-6}/year
Deviation sensitivity	--	150 Hz/Volt
Deviation range	--	±330 Hz
Deviation rate	--	±330 Hz in 25 ms
Linearity	--	±10%
Output waveshape	Sine wave, < 1% distortion	Sine wave, < 1% distortion
Output power	2 mW min. into 50Ω load	1 mW min. into 50Ω load
Spuriour (non-harmonically related)	−60 dB, line-conducted or radiated	−60 dB, line-conducted or radiated
Input power	+15 ±0.3 Vdc, 100 mW max.	+15 ±0.3 Vdc, 100 mW max.

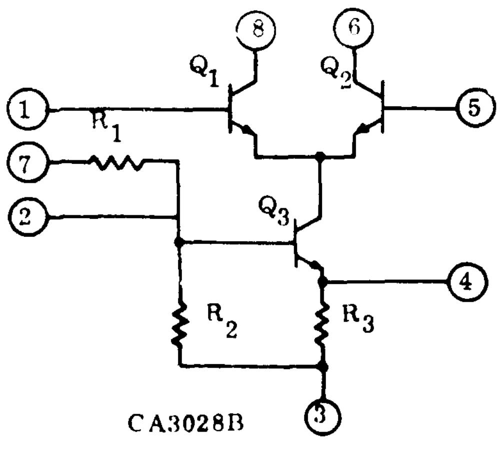
c. Performance of Procured Units. The vendor of the units employed computer analysis to aid in designing the temperatu needed to meet the performance requirements, and all delivered u ment specifications. The drift vs. temperature characteristics of plotted in figures 30 and 31, and the deviation vs. variation in cont sample VCXO is plotted in figure 32.

5. IF AMPLIFIERS AND PHASE DETECTORS.

a. IF Amplifier Circuits. One of the design goals was receiving system at each terminal that would be fully limiting with feedback loop. To achieve this feature without introducing phase s within the dynamic range of the system, the active if. amplifier/li broadband. The noise bandwidth and hence the noise power were c sequentially reduced) by use of passive bandpass filters following e

Figure 33 contains a schematic diagram of the final circuit configu interrogator and transponder recievers, the major portion of the r rowbanding occurs in the circuits of this board. The three-sectior reduces the bandwidth of the first if. signal ahead of the amplifier/ preventing any limiting on noise alone.

RCA CA3028B devices are employed for all active stages. (See eq diagram below.) The amplifier/limiter stages are connected in th where transistor Q3 of the device acts as a constant-current sourc the amplitude at which limiting occurs. The output of each amplifi to the intermediate frequency but is kept relatively wideband (6 to detuning as the stage goes in and out of limiting.



CA3028B

AD-A091 741 CUBIC CORP SAN DIEGO CALIF F/G 17/7
RANGE/RANGE RATE SUBSYSTEM (RRS) FOR COMPLETELY INTEGRATED REFE--ETC(U)
AUG 73 F29601-71-C-0079
UNCLASSIFIED FTR/526-1 AFSWC-TR-73-21 NL

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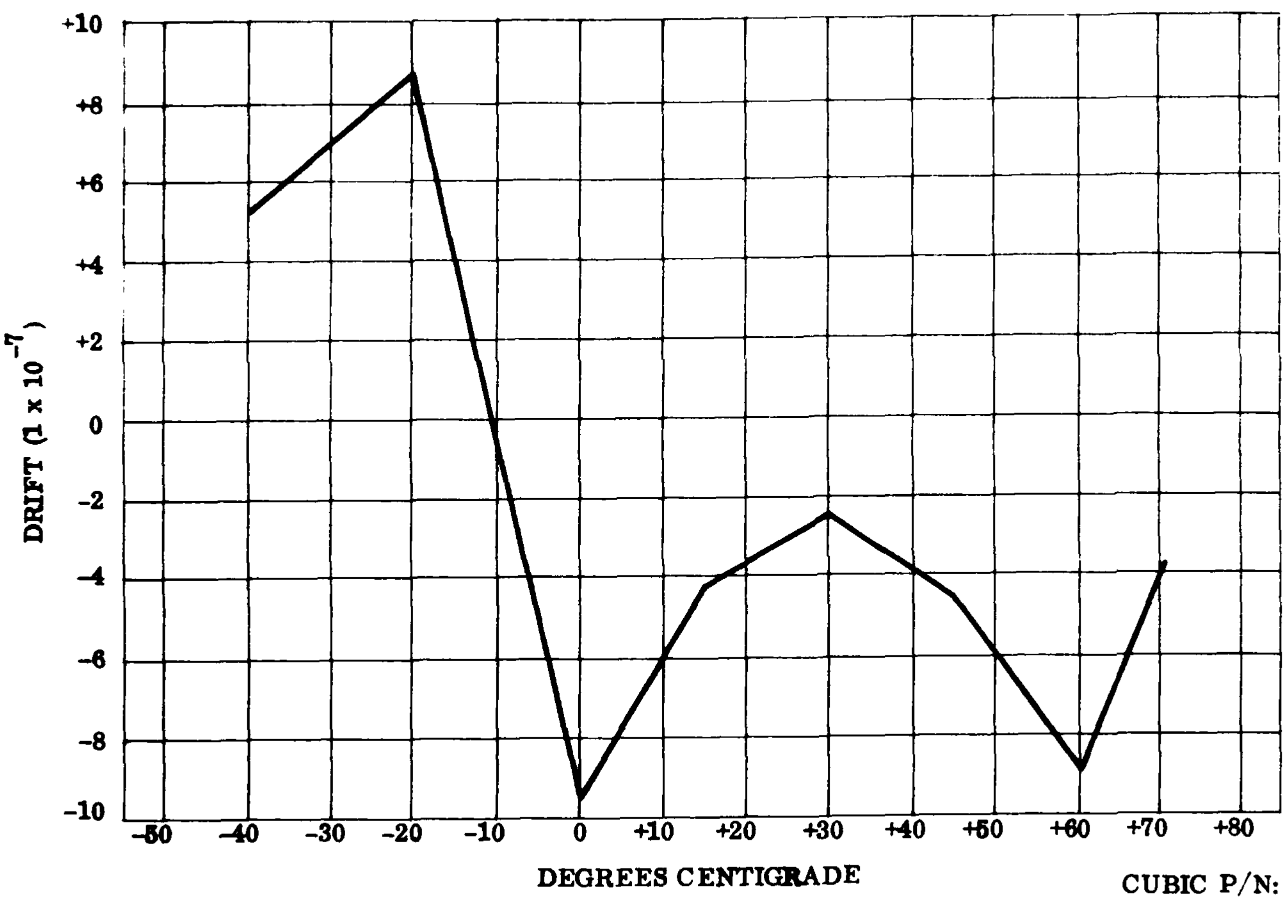
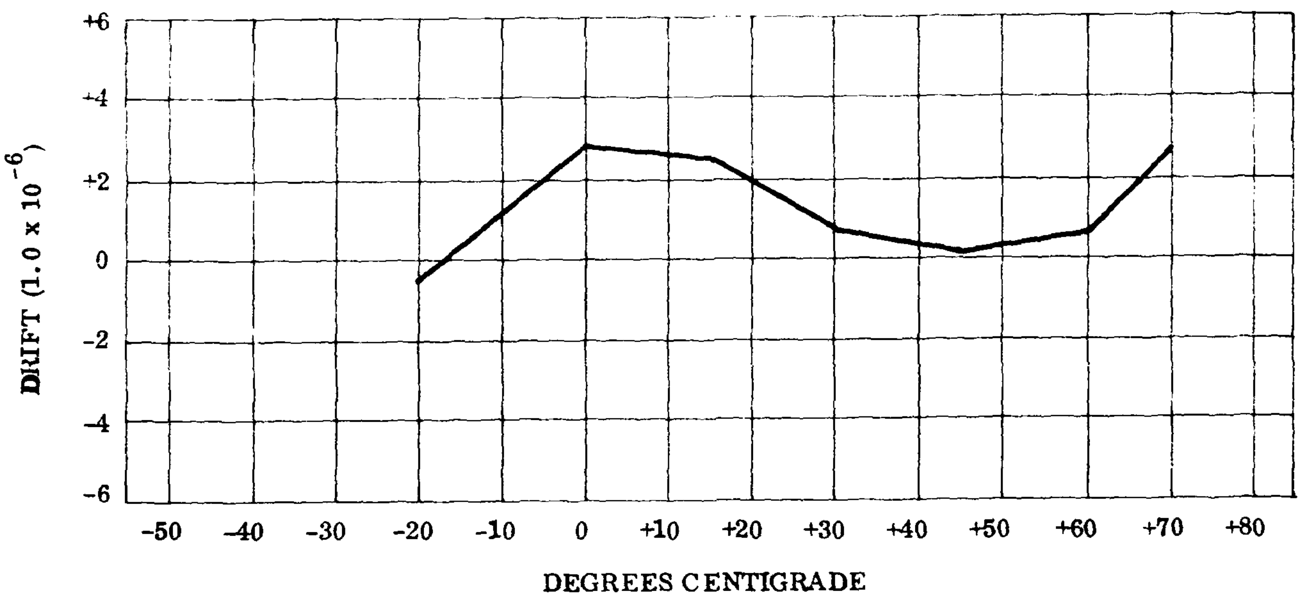


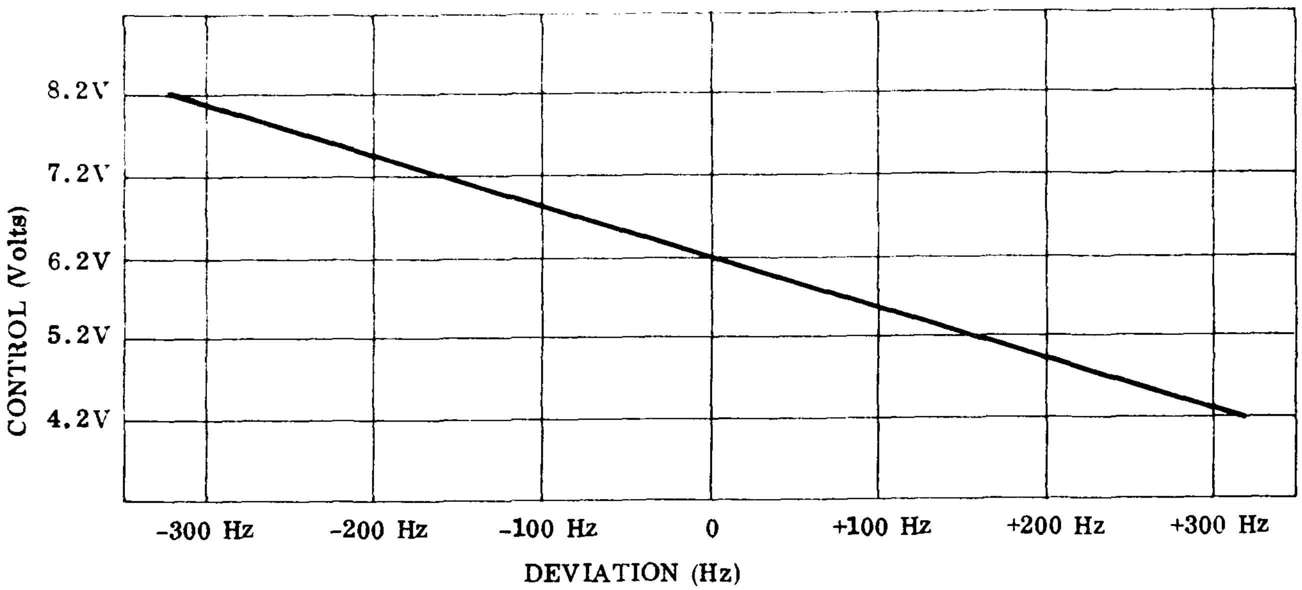
CUBIC P/N: 128327
SERIAL NO: 111E55
CENTER FREQUENCY: 10.866667 MHz
DATE: 30 NOV 71

Figure 30. Frequency Drift vs. Temperature, Crystal Oscillator P/N 128327



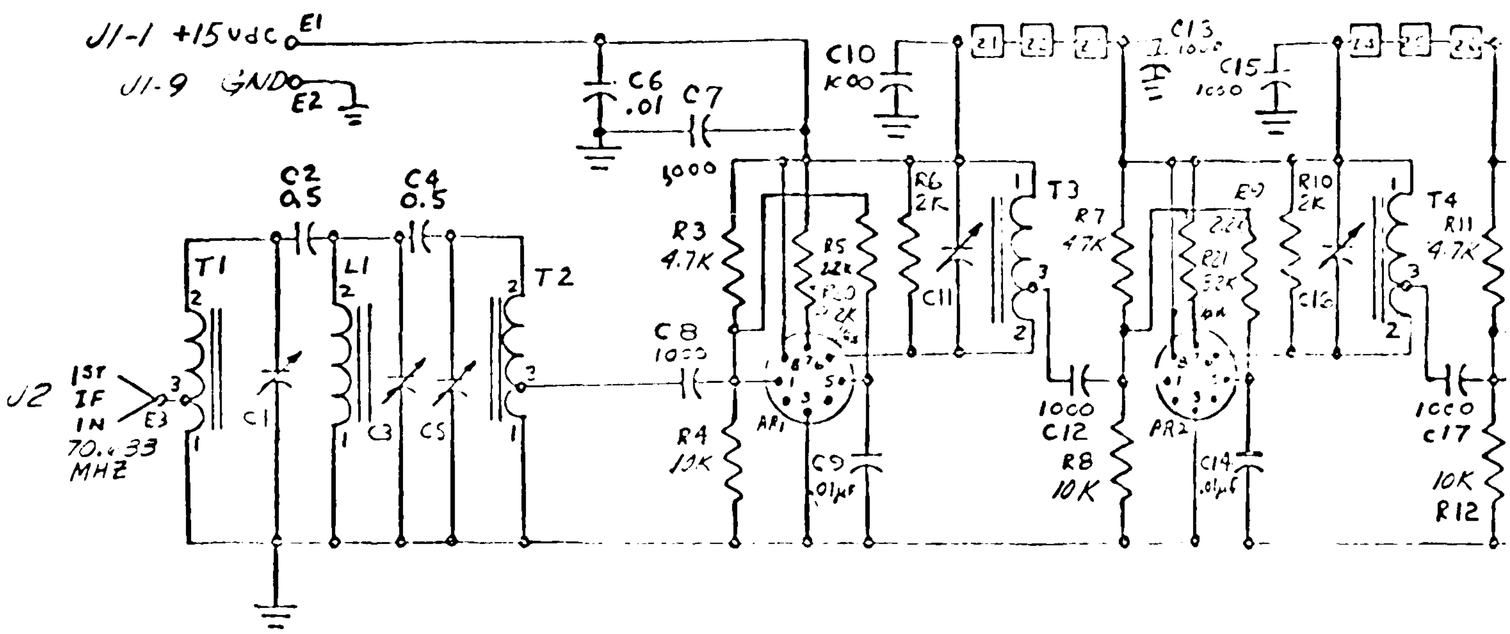
CUBIC P/N: 128334
SERIAL NO: 202EO6
CENTER FREQUENCY: 14.488889 MHz
DATE: 3 FEB 72

Figure 31. Frequency Drift vs. Temperature, VCXO P/N 128334



CUBIC P/N: 128334
SERIAL NO: 202EO6
CENTER FREQUENCY: 14.488889 MHz
DATE: 5 FEB 72

Figure 32. Frequency Deviation vs. Control Voltage, VCXO P/N 128334



3 SECTION BANDPASS FILTER   AMPLIFIER LIMITER   AMPLIFIER LIMITER   AM

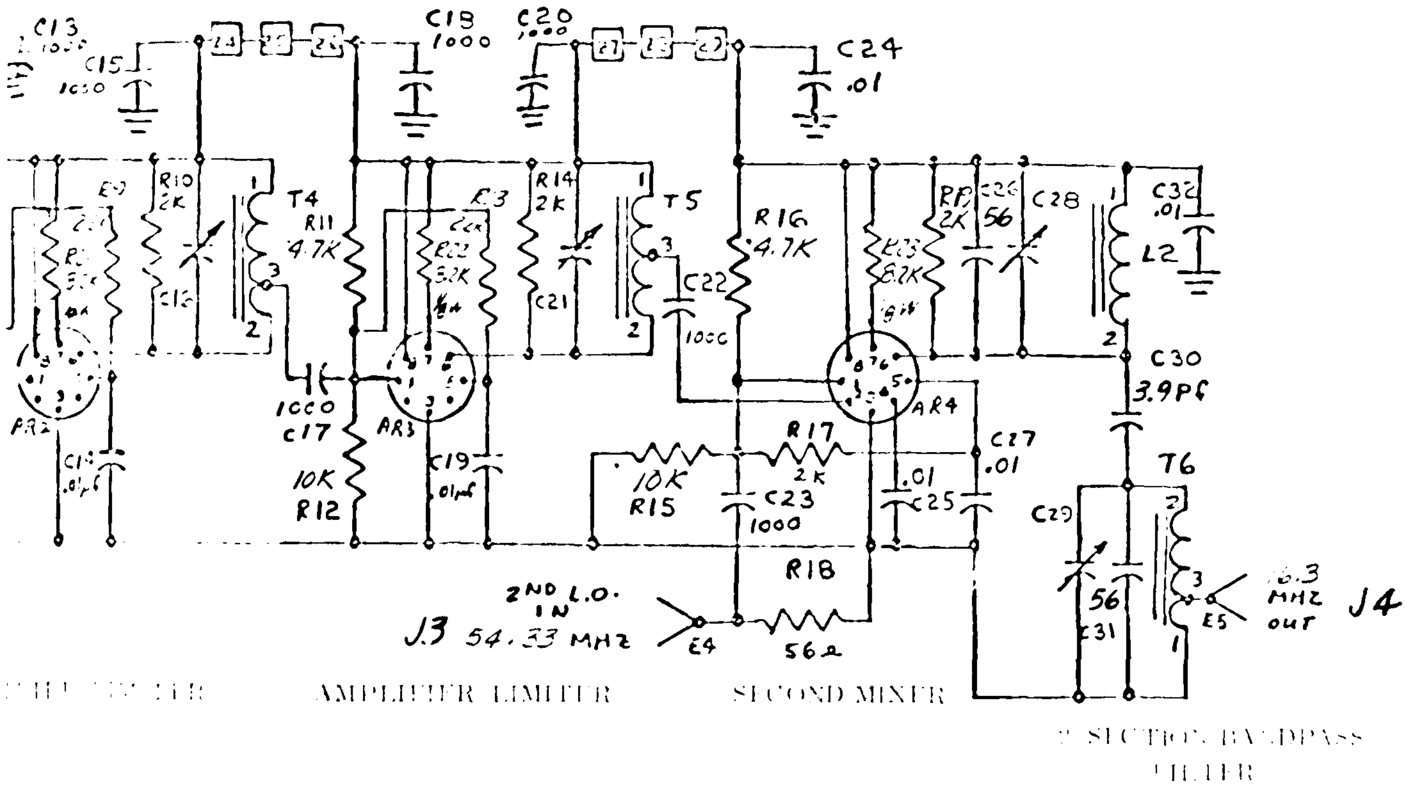


Figure 33. First IF and Second Mixer Circuit Board for Interrogator

For the second mixer, the CA3028B device is connected in the cascode configuration (Q3 and Q2 in cascode) to achieve maximum conversion gain. The second-LO signal then is applied to the differential input (base of Q1). This stage does not operate as a limiter, therefore its output can be narrow-band. The two-section filter at the output of this stage establishes the final predetection bandwidth of the receiver (nominally 1.1 MHz), and the output of this filter is further amplified by stages on the phase detector board described below.

b. Phase Detector Circuits. The phase detector board illustrated in figure 34 contains two additional stages of if. amplification, and also accommodates the two phase detector circuits.

The first if. stage employs another CA3028B device which is connected in cascode for maximum gain. A second CA3028B stage is connected in the differential mode as a limiter, wherein each differential output drives a phase detector. Placing limiters just ahead of the phase detectors ensures that any gain variations in the preceding uncontrolled amplifiers do not introduce signal amplitude variations at the phase detectors.

For the phase detectors themselves, the design goal was to obtain phase detection exhibiting low distortion and minimum offset voltage change over the temperature range from -40 to +71°C and with input signals over the dynamic range of the receiver. A diode quad using matched hot-carrier diodes was found to be the best choice. In each detector, four HP 5082-2800 diodes matched to ±20 mV with forward currents of 0.5 to 5 mA were used. Both the signal reference and if. signals are coupled into the detectors through transformers constructed on high-permeability toroids. The reference signals supplied to the detectors are displaced in phase by 90 degrees, thus one detector extracts phase information and the other obtains amplitude information from the incoming if. signals.

6. TRANSPONDER ID AND VERIFICATION CIRCUITS. For the select-call feature of the RRS, each transponder is assigned a unique 8-bit binary identification number (transponder ID) which, when transmitted as the first element of an interrogation sequence, causes the selected transponder to turn on while all others remain off. To check that the proper transponder is transmitting, a verification feature is incorporated in which the transponder repeats its ID as the first element of its response. PWM/FM/PM techniques are employed at each end, wherein the serial ID data bits are converted to pulse width modulation that frequency-modulates (frequency-shift-keys) a subcarrier oscillator whose output phase-modulates the transmitter carrier. In the initial design concept, the interrogator was to continually transmit the transponder ID until verification was obtained, and the transponder was to continually retransmit its ID until it no longer received ID subcarrier signals from the interrogator. The ID subcarrier signals in both units would then be disabled, and range rate and range measurements could proceed.

a. Statement of the Problem. The initial design of the verification circuitry encountered problems which required a redesign effort in both interrogator and transponder. The predominant problems were associated with receiver noise after the transponder had been selected and turned on, introducing two unsatisfactory conditions:

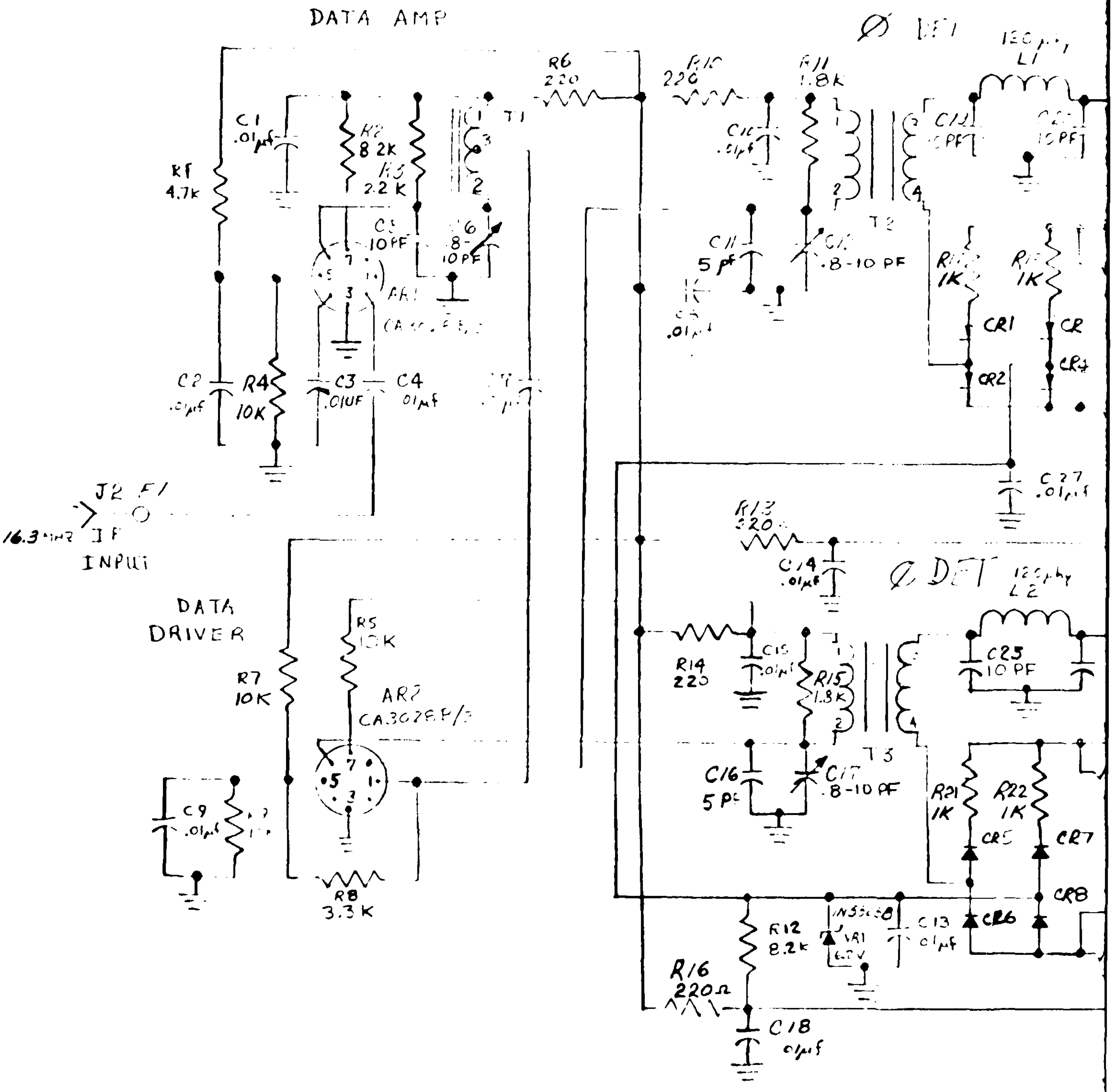
(1) During the ranging portion of the interrogation cycle, noise components in the transponder receiver due to high-index modulation caused false indications of FSK subcarrier lock. Since the transponder verification logic was arranged to respond to detection of the FSK subcarrier, such false lock indications interfered with the proper progression of the interrogation cycle.

(2) During the period in the cycle when the transponder was receiving the FSK data subcarrier and simultaneously retransmitting the verification code, interference and crosstalk developed because of the fact that the transponder uses a low-power portion of its transmitted signal as its first local oscillator. The frequency-shifted verification subcarrier thus imposed on the first LO interfered with the incoming ID subcarrier, causing unreliable operation of the subcarrier lock detector in the transponder and unreliable detection of the verification data at the interrogator.

b. Interim Solution. Subsystem tests demonstrated that selected transponders were reliably being turned on by the select-call method in use, therefore the only system requirement not being satisfied was transponder verification at the interrogator. In discussing this problem with the user it was decided to use the first item equipments without the verification feature so that delivery schedules could be met and system integration plans could proceed. Work was to continue on the problem and ch. nges incorporated later when the solution was found. The verification circuitry therefore was deactivated, and the interrogator circuitry was modified to employ the receiver lock indication (R LOCK) to perform the functions of the verification indication in addition to its usual functions.

c. Final Solution. Early efforts to solve the receiver noise problem included the construction of an input filter and attempts to find a better FSK decoder, neither of which proved satisfactory. It was then decided to eliminate dependence on continuously detecting the FSK subcarrier and to gate off incoming receiver noise during the ranging portion of the interrogation. These modifications were implemented by adding a delay circuit and gate which permit the code to be transmitted for 35 ms after transponder turn-on. A new half-module circuit board was designed and installed in the spare location of the second LO module in the transponder.

The crosstalk interference problem also was solved by eliminating the necessity for continuously detecting the incoming FSK subcarrier after the transponder is turned on. It was determined that, after code match was obtained, continued receipt of the ID code was redundant and thus not necessary. This modification was accomplished by locking in the incoming code after code match, and arranging the interrogator logic



DATA AMP
Ø DET
DATA DRIVER
R6 220
R10 220
R11 1.8K
L1
R1 4.7k
C1 .01µf
R2 8.2K
R3 2.2K
T1
T2
T3
C5 10PF
C6 8-10PF
C11 5 pF
C12 .8-10 PF
AR1
C2 .01µf
R4 10K
C3 .01UF
C4 .01µf
CR1
CR2
CR4
C27 .01µf
J2 E1
16.3 MHZ IF INPUT
R13 220
C14 .01µf
L2
R5 10K
AR2 CA3028A/B
R7 10K
R14 220
C15 .01µf
R15 1.8K
C25 10 PF
C16 5 PF
C17 .8-10 PF
R21 1K
R22 1K
CR5
CR6
CR7
CR8
C9 .01µf
R8 3.3K
R12 8.2k
VR1 6.2V
C13 .01µf
R16 220Ω
C18 .01µf

12-10-??

7

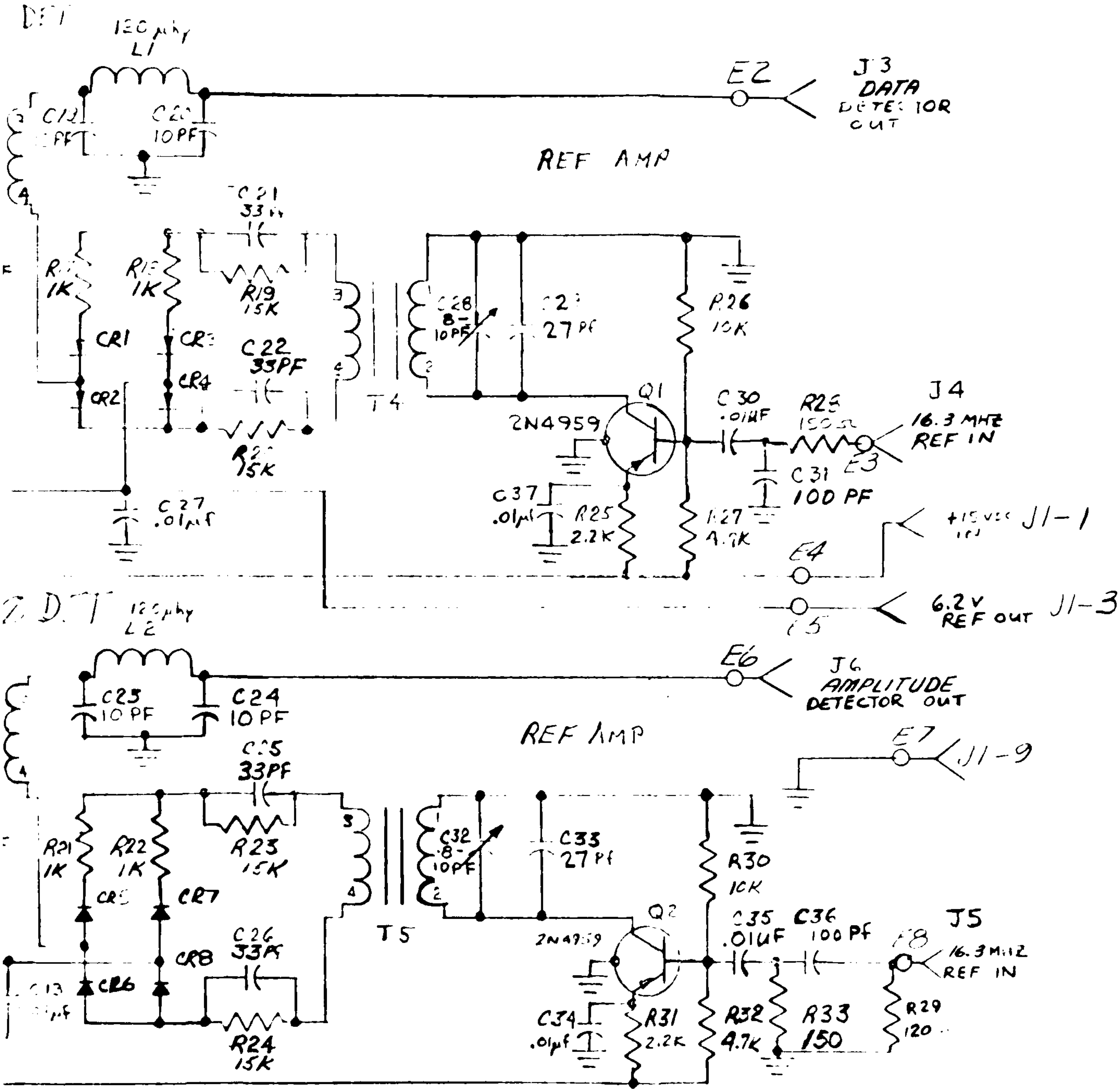


Figure 34. Phase Detector Circui oard

9 9 lank

so that the transmitted ID code shuts off when the interrogator receiver locks to the transponder carrier. Because of the delay circuit that was added, the transponder continues to send its verification data for 35 ms longer, giving the interrogator adequate time to receive and decode it and compare it to the code that was transmitted.

To summarize, in the modified version the interrogator repeatedly transmits the transponder ID until the transponder responds, then discontinues when it detects received carrier lock. The transponder transmits verification data for 35 ms longer, giving the interrogator time to lock to the subcarrier and obtain verification. Following this time-out period, the delayed $\dot{R}$ LOCK signal ramps the range modulation onto the carrier in the usual manner (whether or not an affirmative verification was obtained); and finally, when the data is read out the computer looks at the verification bit in the data quality word to determine if the transponder code was verified.

SECTION VI

EQUIPMENT PACKAGING CONSIDERATIONS

1. INTRODUCTION. Although the circuitry employed in the interrogator and transponder is similar in many respects, their operating environments are quite different, thus the equipment packaging arrangements also differ. For the airborne interrogator, for example, adequate power is available so that equipment cooling can be obtained from a blower, whereas power constraints and the need for sealing against weather conditions require that the ground transponder employ convection/conduction means for thermal dissipation. This section discusses some of the fabrication techniques employed for these two units.

2. AIRBORNE INTERROGATOR. As shown in the photograph in section I, the interrogator is housed in a conventional ATR type case. Overall dimensions of this basic housing are 7.59 in. wide, 7.80 in. high, and 12.56 in. long, with a 3.50-in. extension on the front for the vane-axial blower that supplies the required cooling air. The unit weighs 30 pounds, and mounts in place in the aircraft using the conventional shock-mount base to which it attaches by means of quick-latch hooks at the front and guide pin holes at the rear. Figure 35 shows a side view of the interrogator with the housing removed.

a. Cold Plate Cooling. The primary structure for the interrogator is the cold plate type heat exchanger. The cold plate consists of two plates separated by fins which form air flow channels for the blower-forced air that enters at the front of the unit and exhausts at the rear. The upper and lower plates are 0.062 in. thick, and the fins are 0.040 in. thick. Both plates and fins are constructed of aluminum alloy 6061.

As shown in figure 35, the rf modules attach to the top surface of the cold plate, and the digital circuit boards attach to the bottom surface. Heat is transferred by conduction from the modules to the surfaces of the cold plate, and a Rotron Corp. Model 341JS (Aximax 3) vane-axial fan supplies forced air to the cold plate channels, removing heat to the ultimate sink without circulating the air (which may contain moisture, dust, etc) through the electronics modules. Extensive manual and computer-aided thermal analysis calculations were conducted to ensure that adequare cooling would be obtained at all prescribed altitudes.

b. RF Modules. Figure 36 pictures a typical rf module used in both interrogator and transponder units of the RRS. The basic structure is a cast aluminum housing which accommodates either one large or two small circuit boards contoured to fit into the casting. Covers are provided for both sides for RFI shielding of the signal energy, and a standardized multipin jack J1 interfaces power voltages, control lines and low-frequency signals.

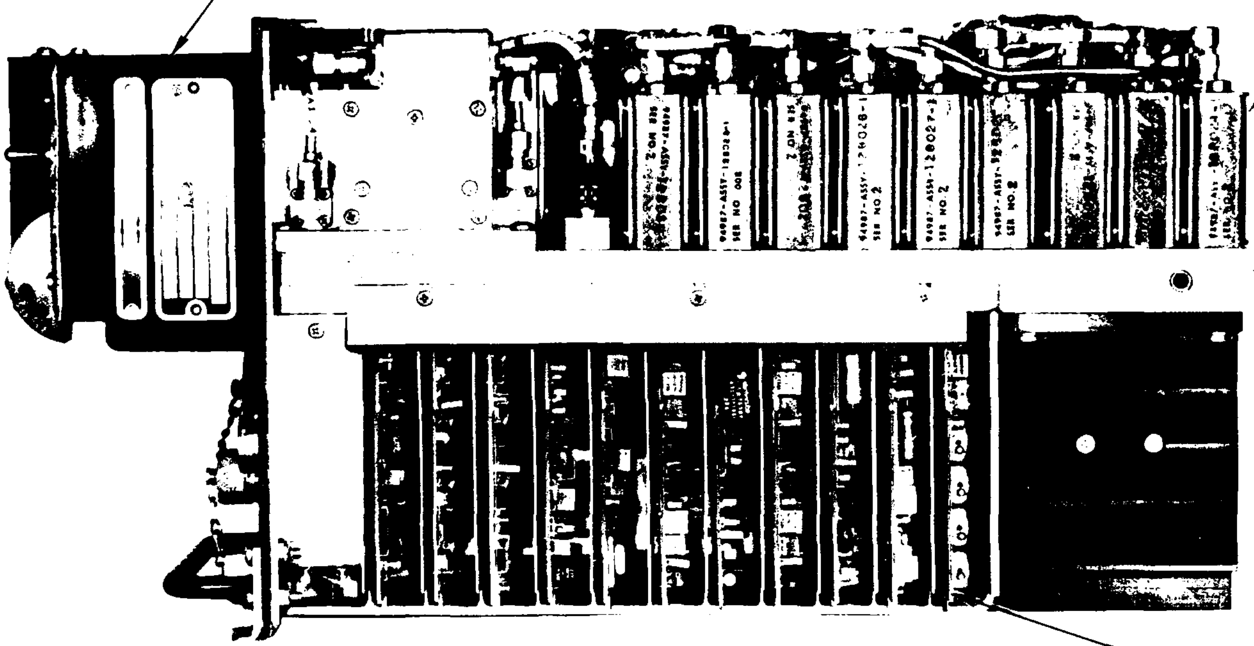


Figure 35. Side View of Airborne Interrogator (Housing Removed)

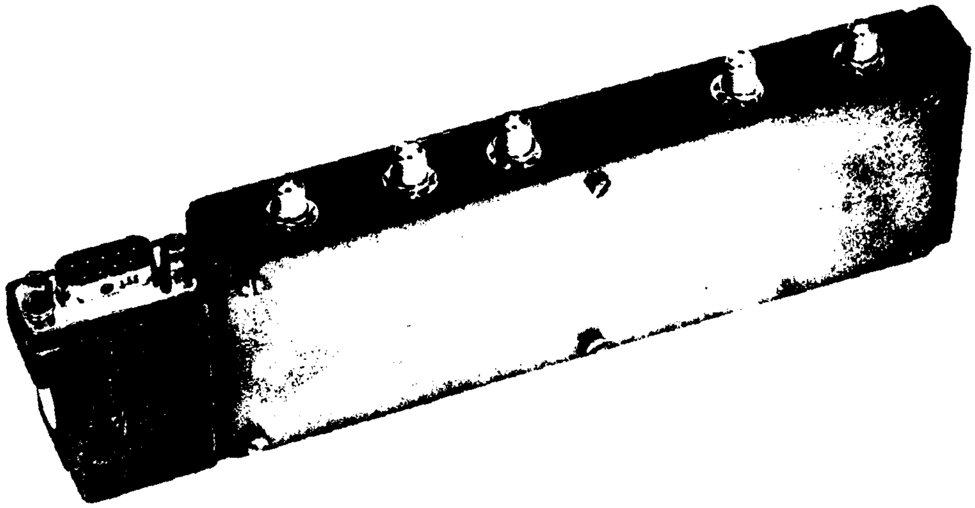


Figure 36. Typical RF Module

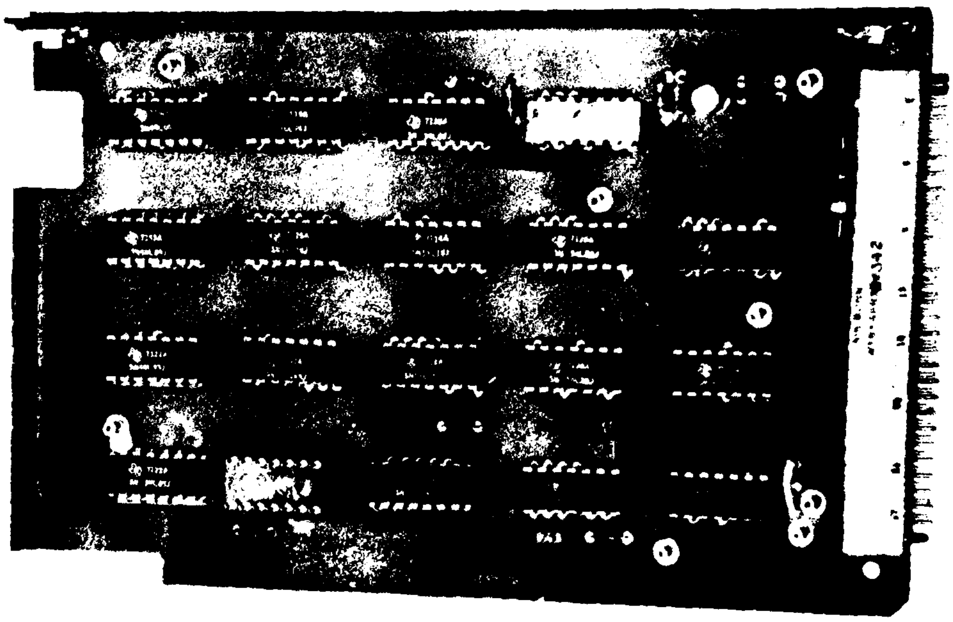


Figure 37. Typical Digital Circuit Board

c. Digital Circuit Boards. Figure 37 shows a typical circuit board for the integrated circuit devices that make up the major portion of the digital logic section of the interrogator (and transponder). These devices mount to a double-sided printed circuit board bonded to an aluminum plate. The plate functions both as a mechanical support and a highly conductive thermal path for enhancing heat dissipation. Heat tests conducted on the first units completed showed that the aluminum plate is not necessary because of the low-power circuitry (less than 1-1/2 watts per cord). Future modules would be built without the aluminum plate.

3. GROUND TRANSPONDER. The ground transponder, also pictured in section I, is packaged in a portable transit case containing the basic transponder, its power supply, and a battery pack. The front panel is provided with a hinged cover containing storage space for cables, voice headset, documents, etc. The entire unit is watertight when the cover is fastened in place. When the cover is open, as for operation in the field, moisture protection again is provided by use of "boots" on the switches and other components on the operating panel.

Figure 38 is a side view of the transponder unit removed from its transit case. Although shown from a different angle, the rf and digital modules employ the same fabrication arrangements as the similar modules in the interrogator.

Heat generated within the transponder is dissipated by a combination of natural convection and radiation. As with the interrogator, detailed thermal analyses were conducted, including laboratory studies using thermocouples installed at various points within the transponder. One such test was conducted in an ambient temperature of 26°C (78.8°F) and a thermal dissipation factor of 55W. Measurements were made at seven different points within the unit after it had been operating under full power for three hours. Extrapolation of all test results to an ambient temperature of 160°F demonstrated that temperatures within the unit will remain within acceptable limits.

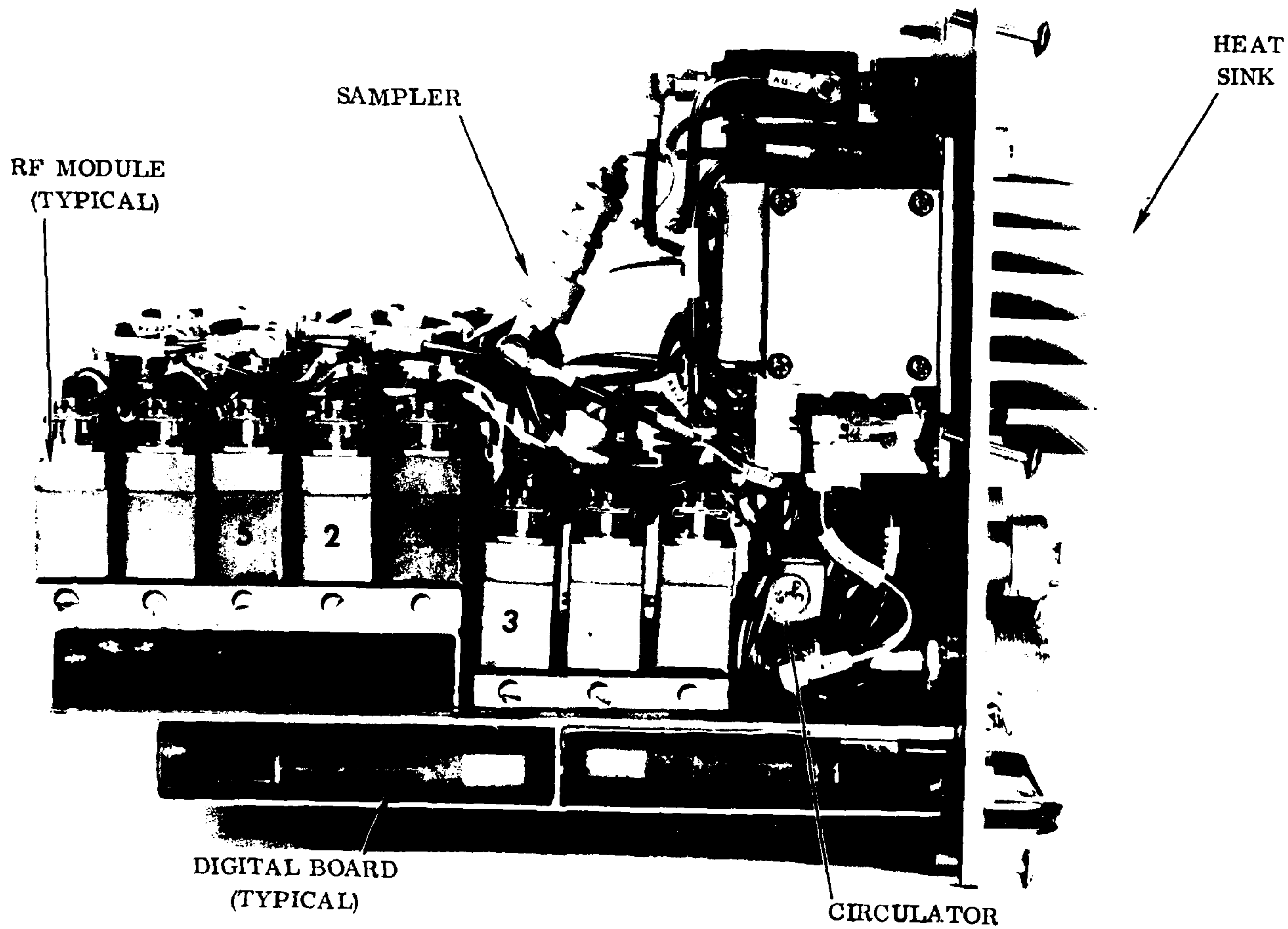


Figure 38. Transponder Removed from Transit Case

SECTION VII

RELIABILITY AND MAINTAINABILITY

1. RELIABILITY ANALYSIS. A reliability analysis was conducted based on the requirements of (a) Reliability Prediction, MIL-STD-756; (b) Reliability Stress and Failure Rate Data for Electronic Equipment, MIL-HDBK-217A, and (c) RADC Reliability Notebook RADC-TR-67-108, Volume II. The analysis demonstrated that the reliability criteria established for the equipment design were met.

For example, for the transponder the design requirement specified a mean time between failure rate (MTBF) of 7,000 hours minimum, allowing a maximum total failure rate of 143 per million hours; the calculated failure rate showed a total of 95 failures per million hours, yielding an MTBF of 10,546 hours. Similarly, the specified minimum MTBF set for the interrogator was 2300 hours (maximum of 435 failures per million hours) whereas the calculated failure rate proved to be 428 failures per million hours, providing a total MTBF of 2336 hours.

The reliability analysis further noted that the bases for the calculated failure rates, as tabulated in MIL-HDBK-217A and RADC-TR-67-108 Vol. II, were established from parts produced during the early 1960's, therefore it can reasonably be predicted that the recently procured parts employed in the RRS equipment will perform more reliably than the calculations indicate.

2. MAINTAINABILITY ANALYSIS. The maintainability analysis, conducted in accordance with MIL-HDBK-472 Procedure II, demonstrated that the maintainability criteria specified in the procurement contract were met. At the organizational maintenance level, for example, the analysis showed that 95 percent of unscheduled maintenance occurances can be corrected within 30 minutes or less (i.e., calculated maintenance time for the transponder: 14.15 minutes; calculated maintenance time for the interrogator: 8.05 minutes). Total maintenance manhours per equipment operating hour were calculated to be 0.0000224 for the transponder, 0.000057 for the interrogator, thus easily bettering the design criterion of 0.01 manhour per equipment operating hour.

SECTION III

CONCLUSIONS AND RECOMMENDATIONS

1. CONCLUSIONS.

a. Equipment Units Produced, Tested and Delivered. A Range and Range Rate Subsystem consisting of one interrogator with antenna, four transponders with antennas and ground planes, and one interrogator test set were designed, fabricated, factory tested and delivered. The operational performance of this subsystem was satisfactory in all respects except for the transponder verification feature discussed in subparagraph b below.

b. Verification Problem Solved. In the initial design of the select call and verification data link, the down-link select call feature satisfactorily turned on the selected transponder but the up-link verification proved unreliable due to receiver noise and interference from the down-link data. To meet delivery schedules the first-item RRS was delivered without the verification capability. Subsequent design modifications solved this data link problem and were incorporated into a follow-on RRS procurement. The first-item subsystem will be retrofitted to correct this problem.

c. New Circuit Designs. Original electronic design work performed in developing the RRS included the following:

(1) A fill-limiting receiver with a large dynamic range (greater than 75 dB);

(2) A new low-noise preamplifier employing microstripline techniques and exhibiting a noise figure less than 4.5 dB;

(3) A new solid-state power amplifier providing output greater than 7.5 watts in the 1.5-GHz frequency region; and

(4) A ground station antenna with lightweight, integral ground plane that folds up into a convenient package, including carrying handle, for easy manpack transport.

d. RRS Oscillators Improved. Very stable fixed-frequency and voltage-controlled oscillators, employing temperature compensation rather than temperature-controlled ovens, are provided in the delivered RRS units. These oscillators represent a product improvement with respect to previous models of the Cubic CR-100 system upon which the CIRIS RRS is based, wherein oscillator drift sometimes caused acquisition problems.

e. RRS Antenna Patterns. Pattern cuts taken on both airborne and ground monopole antennas, when combined to shown system gain, indicate a theoretical possibility of low signal strength under worst-case conditions. However, further analysis shows that, after the CIRIS flight plan is established, appropriately tilting the ground planes for the ground station antennas can minimize or eliminate signal power deficiencies.

f. RRS Reliability and Maintainability. The design of the CIRIS RRS equipment was subjected to reliability and maintainability analyses as specified in the procurement contract. The results, as reported in the Reliability and Maintainability Analysis Report (Cubic Document RMR/526-1 of 2 May 1972), predict that both the reliability and maintainability aspects of the RRS equipment will substantially exceed the specified requirements.

2. RECOMMENDATIONS. Even though the delivered Range/Range Rate Subsystem as presently configured has demonstrated that it will meet all CIRIS operational requirements, the following recommendations are offered which, if adopted, will further enhance operating margins and hence will improve system reliability.

a. Transponder Positioning and Flight Path Geometry. It is recommended that ground transponder sites and the flight path geometry be chosen for best employment of the antenna system gain pattern. When the flight patterns and ground sites have been chosen, the ground planes for the transponder antennas should be tilted to achieve most favorable coverage.

b. Increase in Transmitter Power. Although the transmitter power amplifier for the RRS equipment units is commensurate with the state of the art at the time it was designed, rapid advances in semiconductor technology have now made it possible to provide much higher power for the 1.5-GHz region. For example, Cubic Corporation has developed an experimental alumina substrate power amplifier that yields 20 watts at 1.8 GHz. Modifying the CIRIS RRS to incorporate such a power amplifier would provide greater than 4 dB system improvement; i.e.,

Present power amplifier	7.5W	=	+8.7 dBw
Newly designed amplifier	20W	=	+13.0 dBw
Increase in signal power			4.3 dB

c. Processing Range Data Within RRS Interrogator. To reduce the computation load of the computer and thus save computer time, a logic section could easily be added to the RRS interrogator for combining the four range partials into a binary range word with all ambiguities resolved.

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Final Technical Report

5 AUTHOR(S) (Last name, first name, initial)
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13 ABSTRACT

An equipment procurement comprising a Range/Range Rate Subsystem (RRS) for the Complately Integrated Reference Instrumentation System (CIRIS) is described in detail in a final technical report. An RRS composed of an airborne interrogator and four ground-based transponders is designed, fabricated, factory acceptance-tested, and delivered to the Air Force. With the interrogator operating under the control of the CIRIS airborne computer, and with ground transponders set up for operation at known (surveyed) locations, the RRS measures slant ranges from 200 feet to 200 miles with instrumental accuracies to within 3 feet rms, and measures range rates varying from -5000 to +5000 fps with instrumental accuracies to within 0.03 fps rms under range acceleration conditions from -1000 to +1000 fps^2. The CIRIS computer employs the RRS data to optimally update an inertial measurement unit, providing an accurate, real-time aircraft position, velocity and attitude reference system for aircraft flight test programs.

Predesign analyses of subsystem requirements are discussed, delivered equipment units are pictured and described, and details of the computer/interrogator interface are presented. New circuit developments include a microstripline preamplifier, fully limiting receiver and 7.5W power amplifier, all operating at 1.6 GHz. A transponder verification problem is identified and its solution is provided. The report concludes that the RRS units will reliably meet all specified performance criteria.

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